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by E. J. Habib, G. C. Kronmiller, Jr., P. D. Engels, and H. J. Franks, Jr.

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SUMMARY

A spacecraft tracking system is described which will measure range and range rate for distances varying from 150 to 400,000 km with a resolution of ±15 m in range and ±0.1 m/sec in range rate. This system will provide an order of magnitude increase in tracking accuracy and will use less power at lunar distances than present tracking methods. In addition this system is simple, transportable, and economically feasible.

A preliminary portion of the system, with capability for measuring range only, has been constructed and tested by tracking an aircraft flying over the Blossom Point, Maryland, Minitrack facility; and an analysis of the results shows that the quoted accuracies and performance are feasible. Finally, several system configurations are given as they would apply to specific satellite programs.

CONTENTS

FOREWORD

The system design presented in this report represents the preliminary concepts of Goddard Space Flight Center (GSFC) for a very accurate Range and Range Rate spacecraft tracking system. All aspects of the system embody feasible pieces of hardware, and the overall concept has been flight tested.

A unique feature of the system is the economy, due to the small transponder, in the portion of the satellite required for tracking purposes. The weight of the tracking package in the payload is not expected to exceed 5 pounds, and the average power required for system operation on a duty cycle of 0.05 is under 1 watt. (This is based on the use of 2 watts transmitted power and an overall efficiency of 10 percent.) In all cases, a system bandwidth of 100 cps or less can be realized, as compared with a Radar Range Only System which requires several Mc. As more information on subsystem performance becomes available, the expected system sensitivities can be calculated more exactly.

The present concepts include no contributions from industry, and study and design is continuing at GSFC. It is expected that an excellent system will evolve from this effort.

Because of the early requirement for the Relay I stations, and the progressively later requirements for Synchronous Communications Satellites and the Eccentric Orbiting Geophysical Observatory, the development of the system has been divided into three tasks. The first task is to develop a ranging system for Relay I which excludes ground station modulators, transmitters, antennas and receivers, and the satellite transponder, since these will be provided by the Relay program. Task II will incorporate, in addition, a ground receiver. Task III will provide for development of the modulators, transmitters, antennas, and satellite transponders.

The work reported herein was done in 1961, and represents the state of development of the system at that time.

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INTRODUCTION

With the advent of the artificial satellite as a tool for furthering research in space, the need for precise position determination and prediction on a world-wide basis became a pressing problem. Prior to that time, this need had existed mainly in two fields: The military required knowledge of the position and velocity of an attacking aircraft in order to take effective countermeasures; and the missile age required more exact determinations in order to guide a missile and to evaluate its performance. These situations had in common two salient factors: The distance involved was not too great (in comparison with interplanetary distances), and power for transmitting radio carriers was fairly abundant, although never enough in the case of radar reflection.

Early satellite tracking was characterized by one dominant factor: a lack of sufficient electrical power, first to carry out the intended experiment, and second to provide an adequate communication channel for the data as well as a signal by which the satellite might be tracked. The earliest satellites launched in the United States provided 10 to 20 mw of radiated power for tracking purposes and 0.1 to 0.5 w for the data channel.

Of course, optical tracking provided enough accuracy to satisfy the most stringent requirements; but this was hampered both by insufficient light, either generated or reflected by the satellite, and by the vagaries of the earth's weather. Even so, an optical tracking net was constructed and has operated very successfully.

Successful radio frequency tracking demanded new concepts in the measurement method, and several have evolved. The most successful to date is the application of the interferometer principle to measure the angle between the observer's reference plane and a line connecting the observer and the satellite. *Minitrack* is such a system.

Given a sufficient number of these measurements made at different points along the satellite's orbit, a well designed computer program can produce elements which describe the satellite's past position and velocity with accuracies limited mainly by tracking station abilities, and predict its future positions with accuracies further limited by knowledge of the satellite's environment. Present day abilities can produce satellite angular positions with respect to nearly any position in the world, 24 hours in advance, to accuracies slightly better than 0.05 degree. The need for higher accuracies has not been great until very recently, although it has always existed with respect to orbital mechanics theory, in which higher accuracies are needed to improve knowledge of the earth's gravitational field and the magnitudes of extraterrestrial influence.

To determine which system would best serve the majority of requirements, an appraisal must be made of both the present day systems and those realizable in the near future. A system should be chosen that would: (1) provide an order-of-magnitude increase in tracking accuracy; (2) demand a minimum of carrier power from vehicles as distant as the moon, and further if possible; (3) share the data communication channel; (4) be simple and transportable so that elements of the system could be relocated as needs changed; and (5) be economically feasible.

Tracking systems measure some or all of six basic elements: the three spatial coordinates of a vehicle and the three coordinate rates. Some systems measure pseudo-coordinates. For example, an interferometer measures the x, y, z position of a satellite, but normalized to the range; the rates are inferred by observing the change in x, y, z over a small time interval. A Doppler system measures range rate directly and the second derivative can be inferred.

Two classes of measuring systems exist: those requiring only the emanation of a radio frequency from a satellite, and those requiring responses from the satellite when it is interrogated by a ground-emanated signal. Angle measurements can make use of only the satellite-emanated signal, and the spatial coordinates are determined by combining angle measurements made at several different locations. With prior knowledge of the exact radio frequency, Doppler measurements can be made; and again by combining measurements made at several locations the spatial coordinates may be determined. A range measurement can be made only by measuring the phase delay of a signal or the time of transit of a recognizable modulation of the carrier. This usually requires that the signal originate in the ground station and be transponded back by the satellite. If a signal could be generated on board at a precisely predictable earth time and modulated on a carrier, one-way transmission would suffice for range measurements.

Of these several methods, the Doppler and range techniques are the most easily instrumented; they can yield accuracies limited only by knowledge of the propagating medium and of the velocity of the wave. The interferometer technique, in addition to being hampered by instrumental and environmental conditions, would require frequencies so high that only steerable antennas would provide sufficient sensitivities; and since more antennas would be needed than in a range or Doppler technique, this is ruled out for economic reasons.

Of the range and Doppler techniques many variations are possible. The next section describes what would appear to be a readily obtainable system that will meet all the presently imposed requirements. Parts of this system have been constructed and tested at the Minitrack site at Blossom Point, Maryland, proving the feasibility of the approach taken. A prototype system, to be used mainly for concept evaluation, is now being produced by the Goddard Space Flight Center. Also, ADCOM Inc. is under contract to NASA to provide both a theoretical and experimental basis for choosing many of the system parameters. Of especially great importance is the choice of the modulation method for the ranging portion of the system. Finally, specifications for an operational Range and Range Rate System have been written.

A preliminary analysis of the system sensitivity shows, for example, that with an 85-foot-diameter antenna on the ground and a 5-foot-diameter vehicle antenna in the vicinity of the moon, a transmitter output of only 10 mw is required, and an accuracy (limited only by knowledge of the velocity of light) of 0.4 km is achieved. Replacing the 5-foot-diameter spacecraft antenna with an omni-directional antenna raises the transmitter power output requirement to 1 watt.

A background on the fundamental principles involved and the ultimate accuracies obtainable by measuring range and range rate will be found in Appendix A.

SYSTEM DESCRIPTION

General

The system described here can measure the distance to a satellite from a ground station with a resolution of 15 m and with an accuracy limited only by this resolution, unknowns of the propagating medium, and knowledge of the velocity of light. It can also measure directly the range rate with a resolution of 0.1 m/sec and an accuracy limited only by the aforementioned factors. An ambiguity in range occurs every 937 km; however, this distance can be extended very easily by adding additional lower frequency tones (see Table 1.)

Table 1

Wavelengths for Various Frequencies $\left(\lambda = \frac{299,792.5 \pm .4 \text{ km/sec}}{f}\right)$.

Frequency	Half Wavelength $\lambda/2$		
	(km)	(st. mi)	(ft)
1 Mc	0.1498963	0.093	492
0.5 Mc	0.2997925	0.186	984
100 kc	1.498963	0.931	4,918
20 kc	7.494813	4.657	24,589
4 kc	37.47406	23.285	122,946
800 cps	187.3703	116.42	614,731
160 cps	936.8516	582.13	3,073,654
32 cps	4,684.258	2,910.6	15,368,270
16 cps	9,368.516	5,821	30,736,540
10 cps	14,989.63	9,314	49,178,460
8 cps	18,736.90	11,642.5	61,472,640

The system measures range on the principle that any wave propagated at a given velocity undergoes a delay that depends upon the distance traversed by the wave. It measures velocity on the Doppler principle.

Figure 1 shows the entire block diagram of the range and the range rate system. The system operates as follows: The ground station produces a reference carrier plus sideband pairs obtained by frequency modulating the carrier with several harmonically related lower frequencies. (The lowest of these latter is further divided down to provide both a range and a Doppler reference signal at 10 pulses per second, which is synchronized to WWV and initiates counting of 10 Mc by the Time Interval Units.)

Through digital techniques, the phase delay of the highest frequency signal (100 kc) in going to the satellite and back is measured to an accuracy of 1 percent (15 m) and the remaining lower frequencies (for which one wavelength is greater than the measured range) are used to resolve the ambiguity of how many whole cycle delays have occurred; the combined data constitute the range measurement.

The transmitter carrier is used in performing the range rate measurement; it is derived from the same ultrastable oscillator that provides the sidetone for the range measurement.

The carrier frequency plus sidebands are received in the satellite unit, whose bandwidth can simultaneously handle three ground stations separated approximately by 1 Mc. The transponder unit translates the received signal to a different position in the radio spectrum and acts as a booster amplifier; it has nearly enough total gain to make up for the attenuation of the path from ground to satellite.

On the ground, the translated carrier and sidebands are received in the *range and range rate (RRR) receiver*. Initial mixing with the ground-transmitted reference carrier and phase locking of the receiver to the latter produces a difference frequency that equals the airborne local oscillator's frequency plus its drift and one-way Doppler shift plus the two-way Doppler shift of the carrier. Simultaneously the airborne local oscillator's frequency plus its drift and one-way Doppler shift are received in the *Frequency Translation (FT) Receiver*, where appropriate frequency conversion by means of phase locked oscillators results in an output that is these same frequencies plus a bias frequency. Mixing this output with that of the RRR Receiver removes all local oscillator effects and produces a detected output that is the bias frequency plus the two-way Doppler shift of the carrier.

This detected output is sent to the Range Extraction Unit, which uses narrow band phase-locked loops to separate the ranging frequencies and improve their signal-to-noise ratios. In addition, the outputs are fed to analog phase meters, whose outputs are recorded by an analog paper recorder as a backup.

The 200-kc-biased Doppler is fed to the Doppler Extraction Unit which is set to count 10 Mc in a second Time Interval Unit for a period equal to 6400 periods of the 200 kc ± Doppler frequency. At 1700 Mc, the maximum Doppler expected is ±120 kc. Counting 0.1-μsec units for 6400 periods of 200 kc ± Doppler frequency produces a nonlinear measure of Doppler with an accuracy varying from

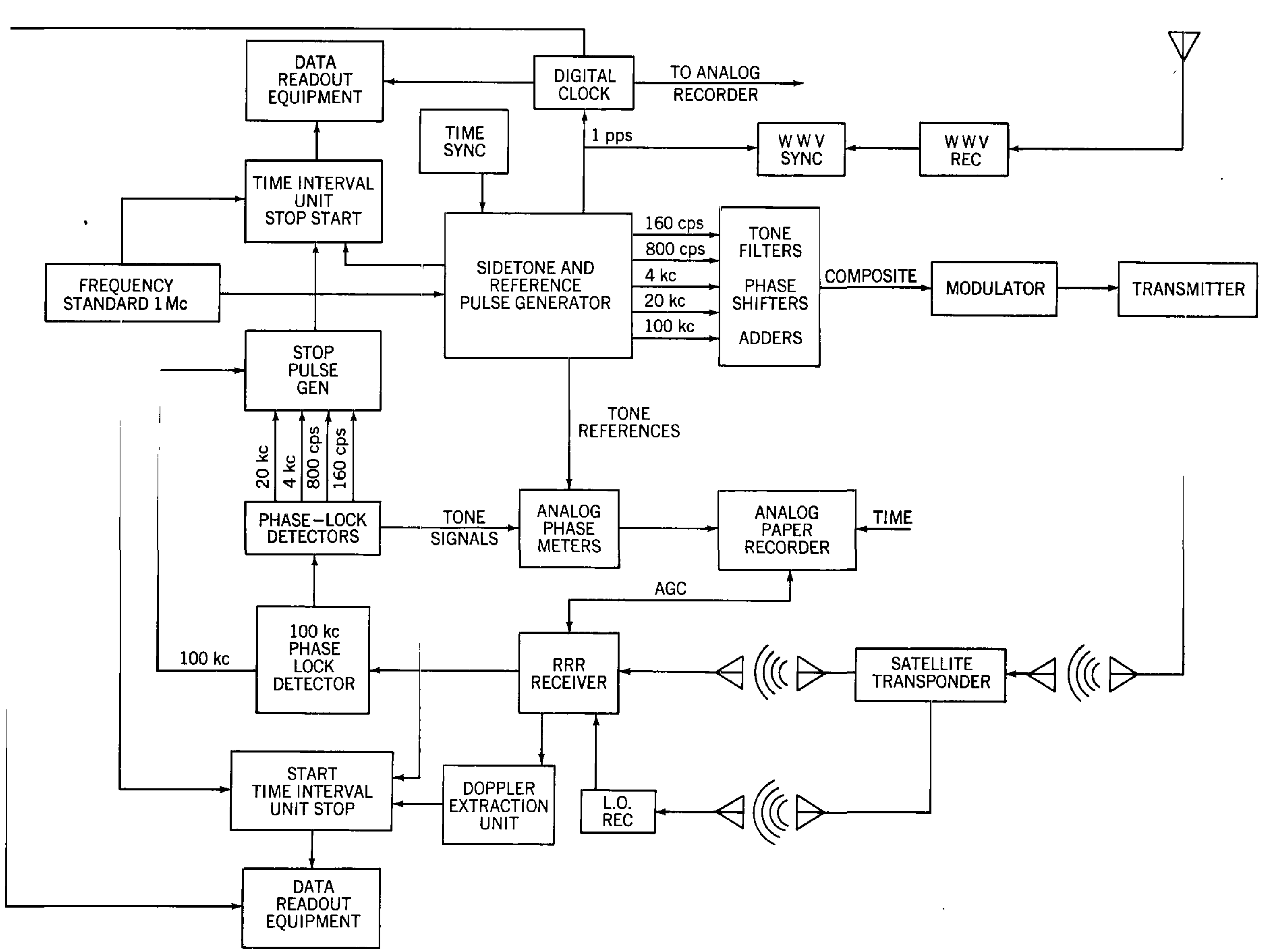


Figure 1—Overall range and range rate system diagram.

approximately 0.25 to 1.0 cps for two-way Doppler, or a nonlinear measure of range rate to an accuracy of approximately 2.2 to 9 cm/sec. The 6400 period counter is gated ON by the 10 pps from the Reference Pulse Generator, thus synchronizing the Doppler Extraction Unit to WWV. The digital outputs from both the Range and the Doppler Time Interval Units are multiplexed with the output from the Digital Clock and punched on two separate teletype punch units, producing data at 10 readings per second in teletype code.

Figure 2 shows a typical trailer layout housing all the necessary equipment for the system. Figure 3 is a typical ground station layout showing the approximate area required and the arrangement of auxiliary equipment for the system.

Sidetone and Reference Pulse Generator

It can be seen in Figure 1 that the heart of the system is the ultrastable (5 parts in 10^{10}) oscillator. As Figure 4 shows, the output of this oscillator feeds the Sidetone and Reference Pulse Generator. The latter, through a digital countdown from the 1 Mc oscillator, produces five harmonically related sidetones: 100 kc, 20 kc, 4 kc, 800 cps and 160 cps. The countdown is continued to form 10 cps as reference start pulses for both the Range and Doppler Time Interval Units (Figure 1) and 1 cps for the Digital Clock. The countdown logic outputs are square waves, so filter and buffer networks are necessary to produce sine waves that can be used in the electro-mechanical phase shifters and for modulating the carrier frequency. The 1 cps sidetone, synchronized to WWV, is used by the Digital Clock to produce days, hours, minutes, and seconds in Greenwich Mean Time for the station.

The phase shifters are used to cancel out any differential or fixed phase shifts that may be present in the various electronic paths. A mechanical linkage with a ratio of 5:1 connects the phase-shifters so that zero-setting of the entire system can be easily accomplished. From the phase shifters the tones feed the Linear Adders and Output Buffer (Figure 5). This section has a switching arrangement that allows for sending any or all

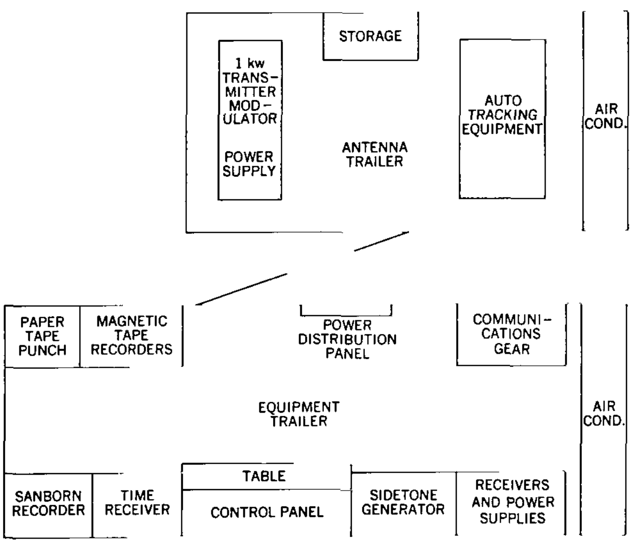


Figure 2—Typical trailer layout.

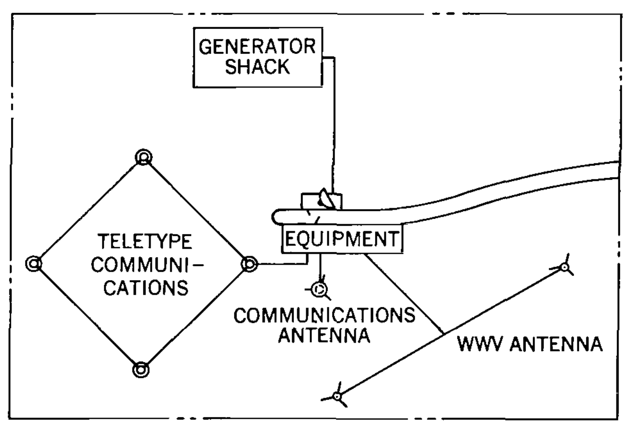


Figure 3—Typical ground station layout.

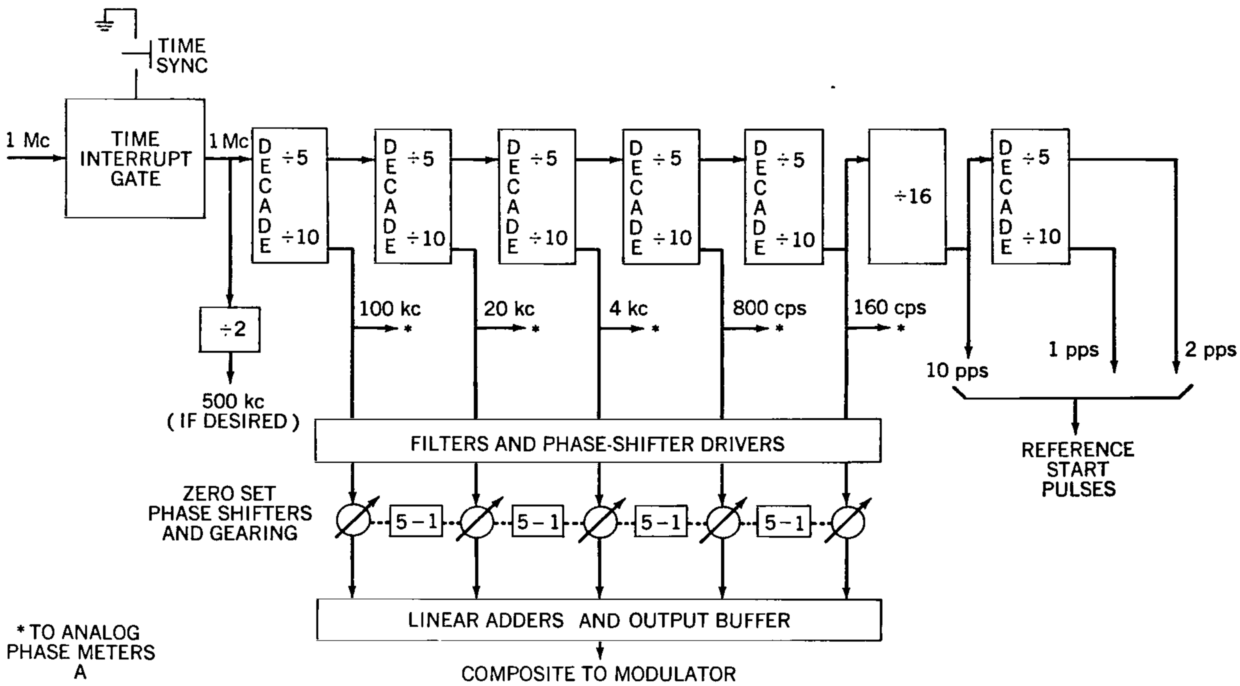


Figure 4—Sidetone and reference pulse generator, tone filters, phase shifters and adders (see Figure 1).

tones with individual amplitude control. Provision is also made for mixing the tones with 100 kc to produce the complementary tones 80, 96, 99.2, and 99.84 kc. This is done in order to gather the FM spectrum so that no components lie too close to the carrier. The composite signal from the Output Buffer then is fed to the FM Modulator and Transmitter (Figure 6). To preserve coherence and to minimize short-term instabilities, the transmitter frequency is derived from the ultrastable oscillator.

Transponder

The satellite borne Transponder (Figure 7) receives the frequency-modulated carrier signal from the ground and mixes the spectrum with a frequency which is (N-1) times the Transponder's basic local oscillator frequency. The relatively low difference frequencies

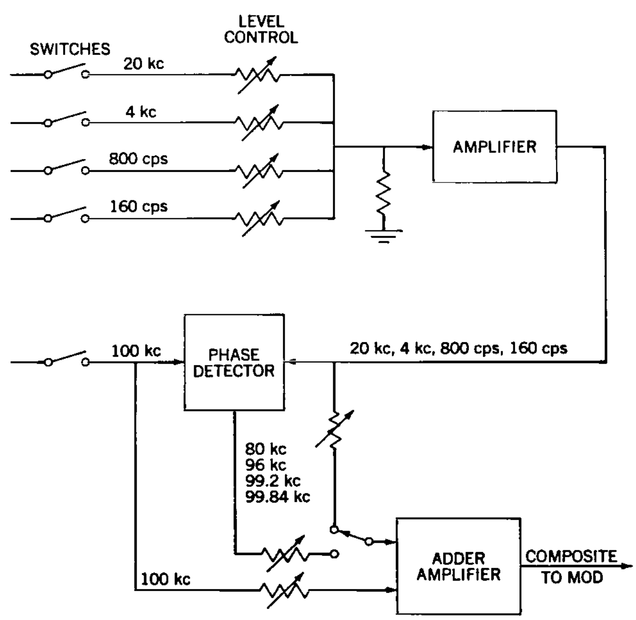


Figure 5—Linear adders and output buffer.

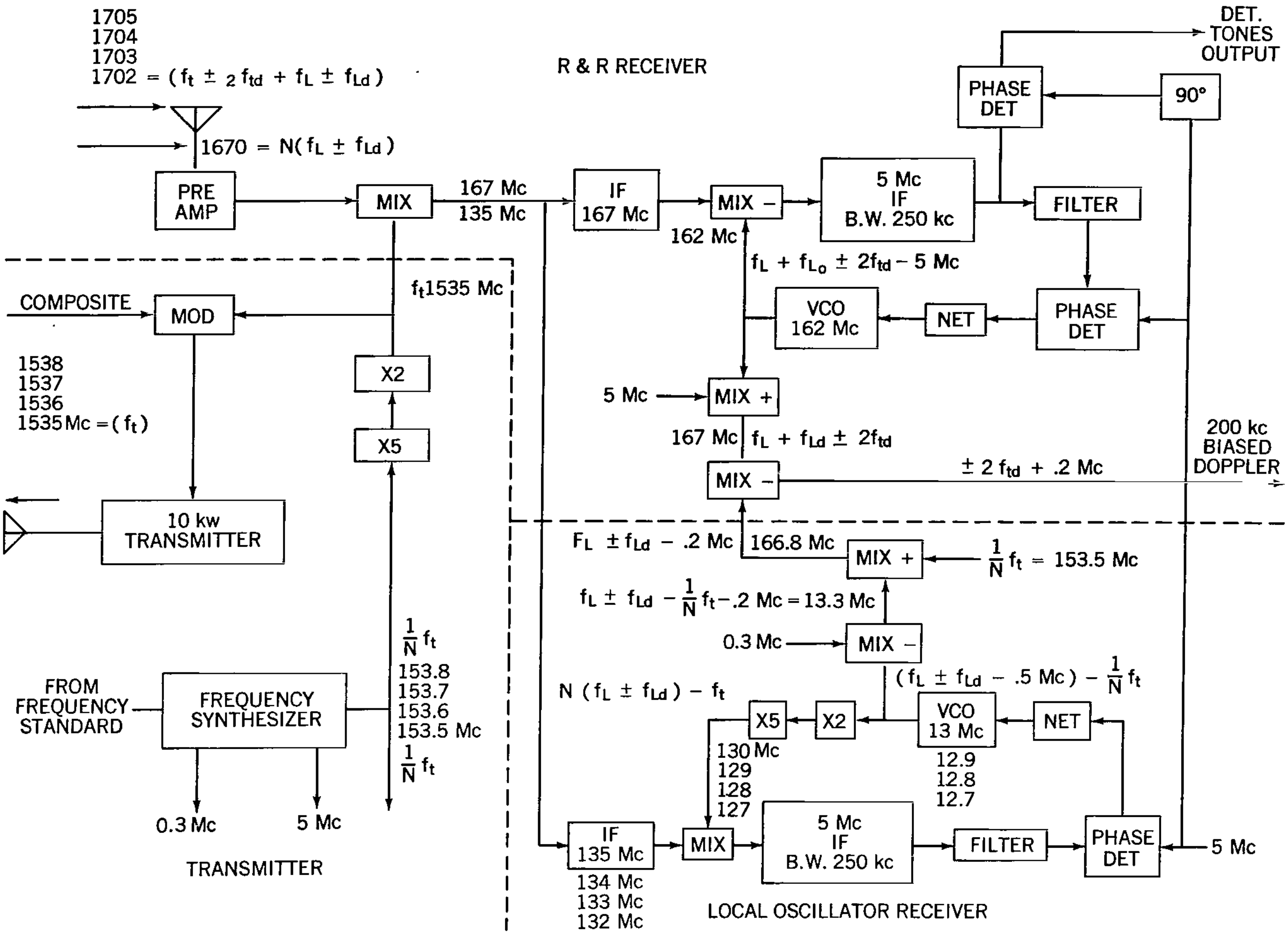


Figure 6—RF ground systems: modulator-transmitter, range and range rate receiver, local oscillator receiver.

(32, 33, 34 or 35 Mc) are extracted and fed to high-gain limiting IF amplifiers. The resulting amplified signal is then converted to a new carrier frequency by mixing with N times the local oscillator frequency. Thus, the received and transmitted carriers differ by exactly the local oscillator frequency. The desired transmitter power is obtained in the power amplifier. In order that the system be coherent, the local oscillator frequency must be available on the ground; thus, N times that frequency is also fed to the power amplifier and transmitted.

Provision is made for three ground stations to operate simultaneously through the transponder, and each receives a coherent translation of its own carrier frequency. It is worthwhile to note that the only other way that coherent transponding could have been accomplished would be to provide each ground station with its own complete transponder in the satellite. A further feature of the transponder is the added 50 kc bandwidth IF strip, for use at extreme ranges where the 100 kc sidetone is not required.

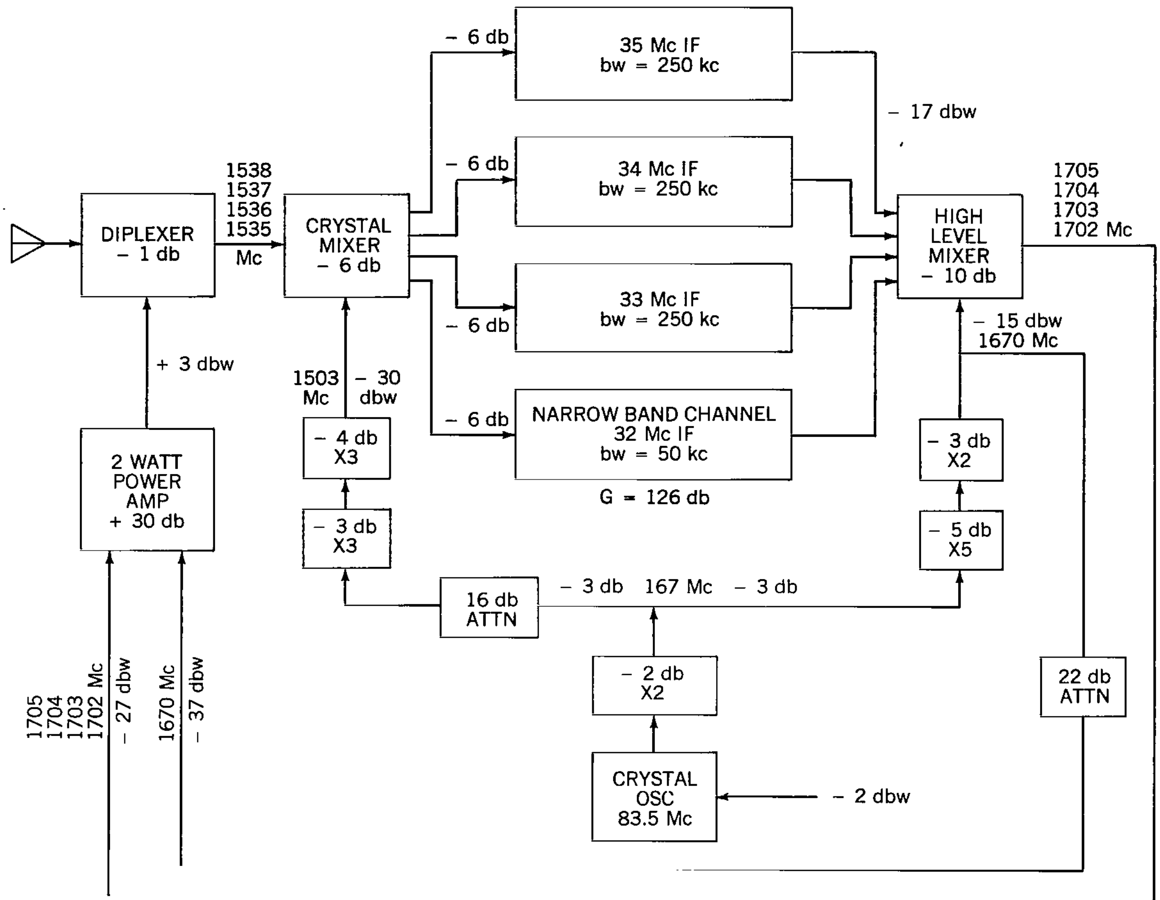


Figure 7—Satellite transponder.

The transponder input contains a diplexer to isolate the transmitted and received signals. The received signals are then jointly mixed in a wide-band crystal mixer with a noise figure of 11 db. The mixing signal is (N-1) times the local oscillator output; thus the mixer output consists of $f_t \pm f_{td} - (N-1) f_L$. Any one of the four IF strips is then selected, depending on f_t. When the output of the IF strip is mixed with N times f_L in a high level mixer, the resulting final output frequency is $f_t \pm f_{td} - (N-1)f_L + Nf_L$, or $f_t \pm f_{td} + f_L$, which is then amplified by the power amplifier and retransmitted. At the same time, the local oscillator frequency is multiplied by N to produce Nf_L, which is also transmitted to the ground station.

Ground Receiver and Signal Detection

The transponded signal is received by the range and range rate receiver shown in Figure 6. After preamplification in a low-noise-figure preamplifier, the signals are mixed with the ground transmitter frequency and converted to the first IF frequency. Two IF's are necessary: one for the sidetone signals and one for the transponder local oscillator frequency. Synchronous detection is used to regain the sidetones, and the transponder local oscillator frequency is used to remove the effects of its own Doppler and drift. An offset frequency of 200 kc is introduced to facilitate the Doppler counting by preventing it from passing through zero.

Referring again to Figure 6, it is noted that one of the received frequencies, "A", nominally 1702 Mc, can be thought of as the transmitted frequency (f_t) ± twice its Doppler shift $(2f_{td})$ + the transponder local oscillator frequency (f_L) ± its Doppler shift (f_{Ld}). The other, "B", nominally 1670 Mc, can be thought of as $N(f_L \pm f_{Ld})$. The frequency of the first local oscillator of the receiver, f_t, is nominally 1535 Mc. Thus, from "A" we obtain

$$f_t \pm 2f_{td} + f_L \pm f_{Ld} - f_t = \pm 2f_{td} + f_L \pm f_{Ld} ,$$

$$= 1702 - 1535 = 167 \text{ Mc}$$

which we call A'. From B we obtain

$$N(f_L \pm f_{Ld}) - f_t$$

$$\text{or} \quad f_t - N(f_L \pm f_{Ld}) = 1670 - 1535 = 135 \text{ Mc } (N = 10) .$$

Let this equal B'. The frequencies A' and B' are the first IF frequencies. Since these frequencies are very high, the first IF frequencies need only enough gain to offset the loss from mixing and to feed the second mixer. In the second mixer A' is mixed with a frequency from a phase-locked oscillator that differs from A' by 5 Mc. The result is

$$f_L \pm f_{Ld} \pm 2f_{td} - (f_L \pm f_{Ld} \pm 2f_{td} - 5 \text{ Mc}) = 5 \text{ Mc} ,$$

which is the second IF frequency. A second phase detector is used to separate the sidetones from the 5 Mc IF.

Frequency B' is also mixed with a frequency from a phase-locked oscillator that differs from it by 5 Mc. Thus we have (for $N = 10$)

$$-f_t + N f_L \pm f_{Ld} - N\left(f_L \pm f_{Ld} - 0.5 \text{ Mc} - \frac{1}{N} f_t\right) = 5 \text{ Mc}$$

since in this voltage controlled oscillator $f_L \pm f_{Ld} - 0.5 \text{ Mc} - f_t/N$ was also generated.

To obtain the Doppler information, we take $f_L \pm f_{Ld} - 0.5 \text{ Mc} - f_t/N$ and mix it with 0.3 Mc instead of 0.5 Mc to provide the 200 kc bias frequency and then mix with f_t/N to obtain $f_L \pm f_{Ld} - 0.2 \text{ Mc}$.

In section A of the receiver we mix 5 Mc with $\pm f_{Ld} + f_L \pm 2f_{td} - 5 \text{ Mc}$ to form $\pm f_{Ld} + f_L \pm 2f_{td}$. Next this is mixed with $f_L \pm f_{Ld} - 0.2 \text{ Mc}$ to obtain $\pm 2f_{td} + 0.2 \text{ Mc}$, which is then fed to the Doppler extraction unit.

Range and Doppler Extraction Units

The tones from the receiver may be either complemented or uncomplemented; therefore in the Range Extraction Unit (Figure 8) an additional phase detector may be necessary, for the complemented

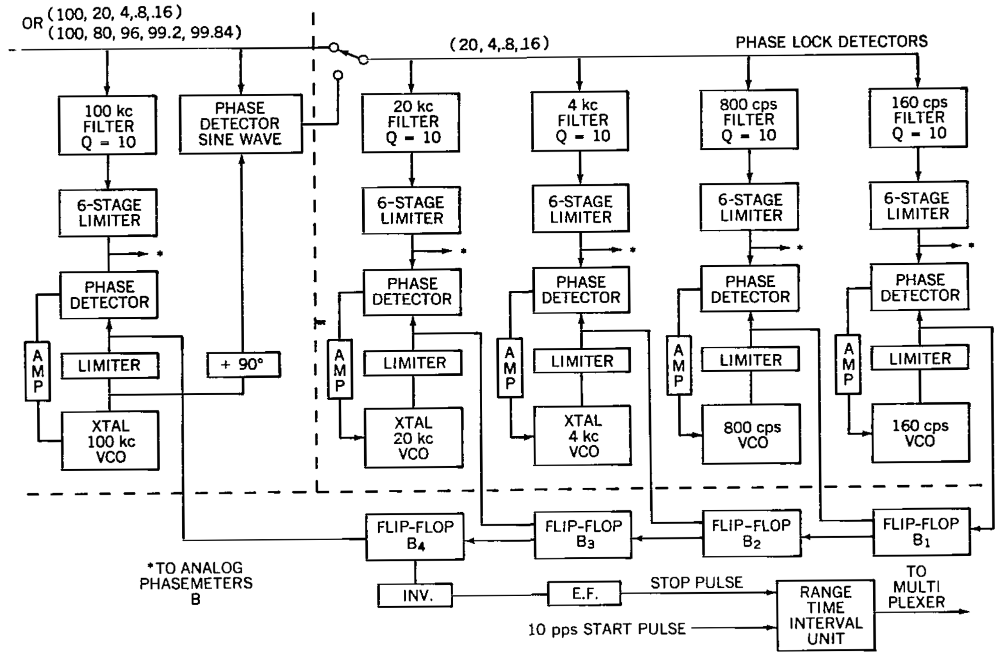


Figure 8--Range extraction unit.

case, to retrieve the original sidetones. The tones are separated by bandpass filters and stable voltage-controlled oscillators are locked to them, thus making possible the use of very narrow loop bandwidths.

The tones, when properly phased, are ready for the stop pulse generator. The proper phasing is shown in Figure 9 along with the action of the flip-flop circuits. The positive-going edge of the 160 cps tone turns flip-flop B_1 "on", allowing the next positive rise of 800 cps to turn it "off" and thus send an "on" pulse to B_2. This operation continues through the higher frequencies until the 20 kc wave selects the proper 100 kc wave and a stop pulse is sent to the counter. This method of ambiguity resolving allows for up to ±10 percent phase noise on each tone before ambiguities cannot be resolved. The output of this logic is a single pulse at a repetition rate equal to the lowest frequency (160 cps), and a rise time definition of 1/2 percent of the period of the highest frequency (100 kc). The proper phasing of the tones is accomplished by rotating the individual phase-shifters, and the delay between start time and stop time for zero range is removed by employing a simulated transponder to rotate the phase-shifters as a unit. The actual range measurement is made in the Range Time Interval Unit which is started by the 10 pps start pulse and counts at a 10 Mc rate until stopped by the stop pulse.

As an auxiliary measure, analog phasemeters and analog recording are provided as shown in Figure 10. Reference tones are obtained from the tone generator and signal tones from the Range Extraction Unit. Receiver AGC and time signals are also displayed on the recorder.

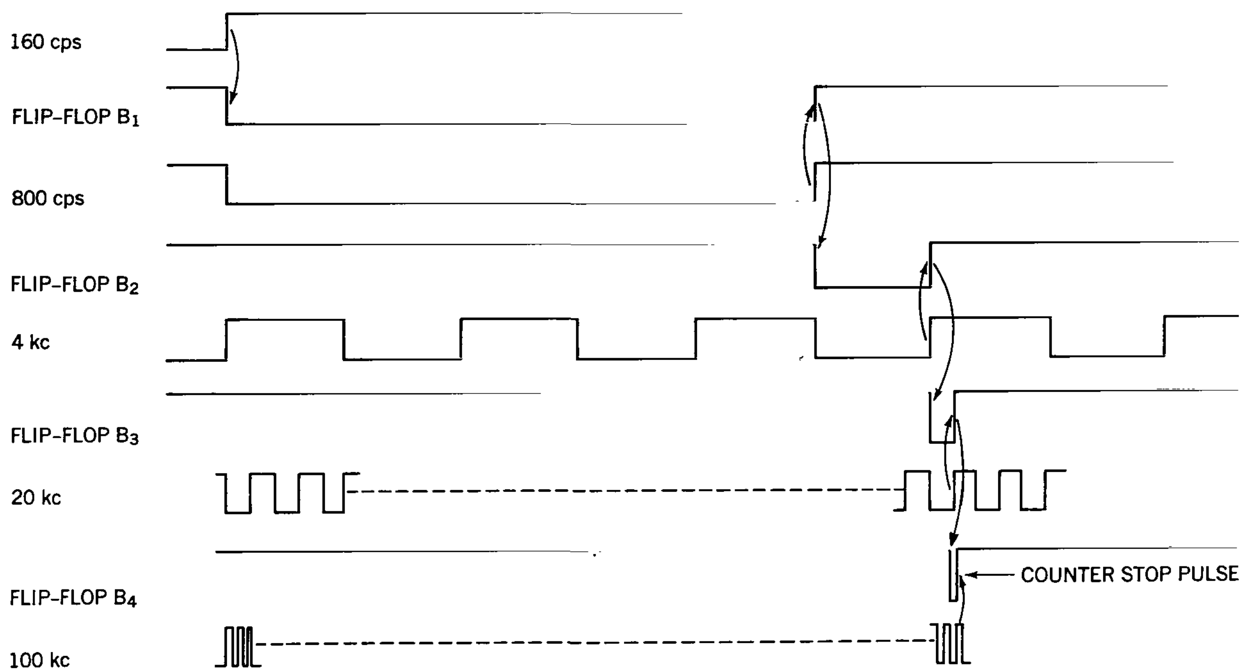


Figure 9—Sidetone phasing diagram.

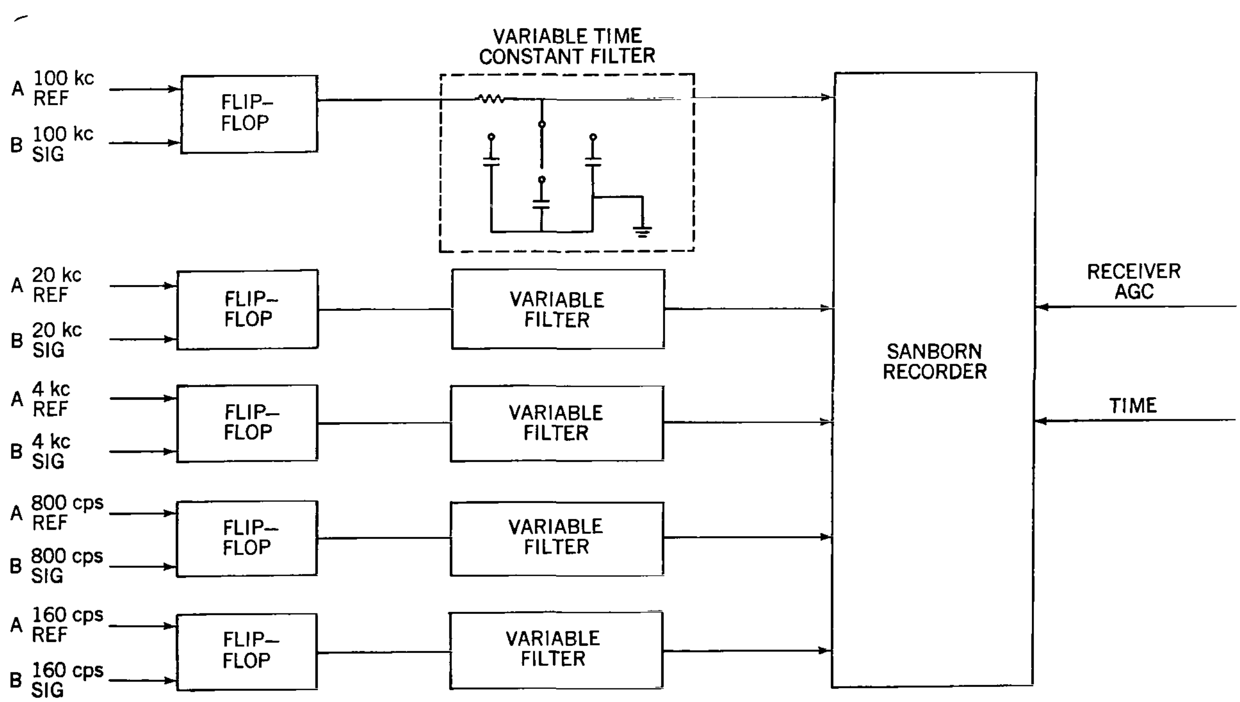


Figure 10—Analog phasemeters.

Doppler information is obtained from the Doppler Extraction Unit (Figure 11). The Doppler signal with its 200 kc offset is gated by the 10 pps start pulse and fed to a preset counter which starts and stops the time interval unit. The time interval unit counts at a 10 Mc rate.

The range and Doppler information, together with time from the Digital Clock, is sent to the multiplexer and buffer storage for printout on paper tape punches as in Figure 12. The punches then present the information ready for teletype relay to the computing center.

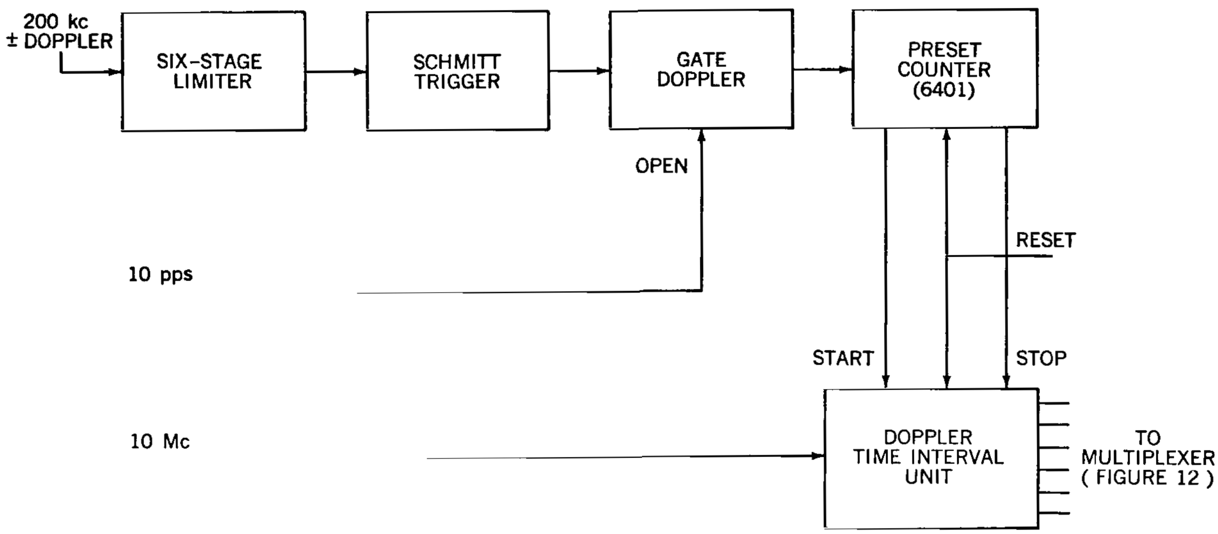


Figure 11—Doppler extraction unit and time interval unit (see Figure 1).

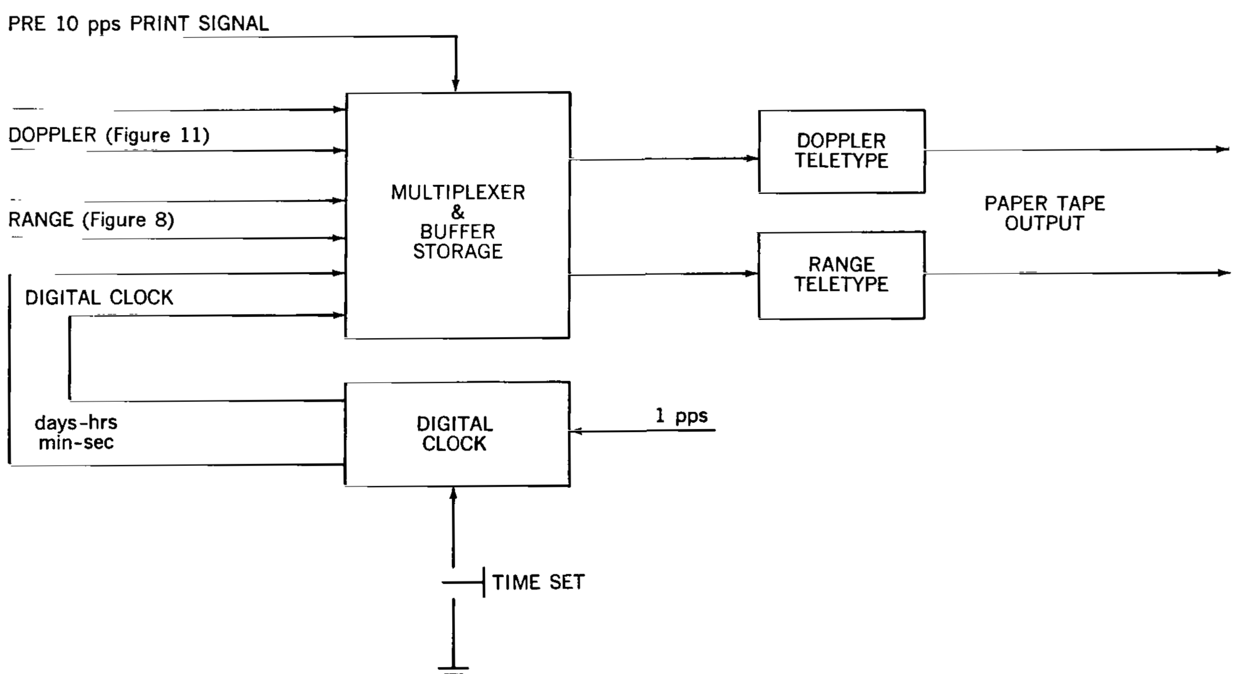


Figure 12—Data readout equipment and digital clock (see Figure 1).

Antennas

The choice of antennas for the ranging system is essentially determined by two categories of operation. The first is for satellite altitudes below 10,000 km and the second for altitudes beyond 10,000 km. To track satellites up to 10,000 km, high tracking rates will be required; and since the system will be used to produce either guiding or predicting information for very narrow beamwidth antennas (less than 1 degree) a beamwidth of at least 3 degrees will be required. Because this is still a very narrow beamwidth, autotracking is required.

With these considerations in mind, a 14 ft diameter dish antenna with a gain of 35 db at 1600 Mc, or its equivalent, has been chosen. On the basis of this antenna and a range of 10,000 km, the minimum system is calculated as follows:

When the range is beyond 10,000 km, the higher sidetones are not necessary and a reduction in transponder bandwidth can be realized. Thus, for a range of 100,000 km, the minimum system calculation is as shown in Tables 4 and 5.

These calculations represent the minimum satellite transmitter power required. They are very conservative as no satellite antenna gain is assumed, the ground receiver noise figure of 3 db can be bettered and the required signal-to-noise ratios have a 6 db margin included. Also, very conservative receiver and sidetone bandwidths are assumed. They may be bettered by approximately 10 db.

For transmission to the moon (400,000 km) a vehicle antenna gain of 25 db is assumed and an 85 ft diameter ground antenna is mandatory.

The distance is about four times (12 db) greater than the 100,000 km case; therefore, the net gain is +13 db (Tables 6 and 7).

Table 2

Characteristics of the Minimum System for Transmission to Satellite at 10,000 km.

System Parameter	Value
1535 Mc ground transmitter power	+60 dbm
Ground antenna gain	+35 db
Polarization loss	-3 db
Path loss (1535 Mc)	-176.5 db
Vehicle antenna gain	0 db
Received total power (satellite)	-84.5 dbm
Receiver noise power at 250 kc bandwidth and noise figure of 11 db	-109 dbm
Total signal-to-noise ratio in vehicle	+24.5 db

Table 3

Characteristics of the Minimum System for Transmission to Ground at 10,000 km.

System Parameter	Value
Vehicle antenna gain	0 db
Ground antenna gain	+35 db
Polarization losses	-3 db
Path loss (1705 Mc)	-177.3 db
Total losses	-145 db
Receiver noise power at 100 cps bandwidth and noise figure of +3 db	-152 dbm
Required carrier power for carrier-to-noise ratio of +20 db	-132 dbm
Required transmitter carrier power	+13 dbm
Required minimum total transmitter power for remaining 5 sidetones at 10 cps bandwidth each and S/N = +30 db (each sidetone requires -132 dbm) add 7 db	+20 dbm
Local oscillator leak power	+13 dbm
Total transmission power (13 dbm + 7 db + 3 db) 200 milliwatts	+23 dbm

Table 4

Characteristics of the Minimum System for Transmission to Satellite at 100,000 km.

System Parameter	Value
Transmitter power increase 1 kw to 10 kw	+70 dbm
Ground antenna gain	+35 db
Polarization loss	-3 db
Path loss (1535 Mc)	-196.5 db
Vehicle antenna gain	0 db
Received total power	-94.5 dbm
Receiver noise power with a bw of 40 kc and a noise figure of 11 db	-117 dbm
Total signal to noise ratio in vehicle	+22.5 db
Additional ground antenna gain by using 85-foot diameter antenna (total gain = 50 db)	+15 db
Total signal-to-noise ratio in vehicle	+37.5 db

Table 5

Characteristics of the Minimum System for Transmission to Ground at 100,000 km.

System Parameter	Value
Vehicle antenna gain	0 db
Ground antenna gain	+35 db
Polarization losses	-3 db
Path loss (1705 Mc)	-197 db
Total loss	-165 db
Receiver noise power at 30 cps band-width and a noise figure of 3 db	-156 dbm
Required carrier power for carrier-to-noise ratio of +20 db	-136 dbm
Required minimum carrier power	+29 dbm
Required power for sidetones	+7 dbm
Required power for local oscillator	+3 dbm
Total transmitter power (minimum)	+39 dbm (10 watts)
85 ft diameter antenna gain	+50 db
Total transmitter power required	+24 dbm (250 milli-watts)

Table 6

Received Signal to Noise Ratio for Transmission to Satellite at 400,000 km.

Vehicle Antenna Gain (db)	Received Signal-to-Noise Ratio (db)
+25	+50.5
+7	+32.5

Table 7

Required Transmitter Power for Transmission to Ground at 400,000 km.

Vehicle Antenna Gain (db)	Required Transmitter Power
+25	+11 dbm (+12 milliwatts)
+7	+28 dbm (0.63 watt)

RANGING SYSTEM APPLICATIONS

At present it is contemplated that the ranging system will be required on the Relay I, the Synchronous Communications (SYNCOM) and the Eccentric Orbiting Geophysical Observatory (EGO) satellites and for Atlas Agena injection tracking.

Relay I Satellite

The Relay I communications satellite has a planned eccentric orbit with an apogee of 4000 km. Because the satellite's flight equipment had been established, the only system which could be used with this satellite was one that employed the existing electronic equipment on board. This limitation made it impossible to incorporate both a Range and a Range Rate System. A Range Only System, however, could be used without modification of the existing on-board transponder, and by using only a part of the ground based ranging equipment. Figure 13 shows the block diagram of the equipment provided. The existing Relay I receivers, transmitters, antennas, and transponder were used on a time-share basis with the communication experiments which constitute the satellite's mission.

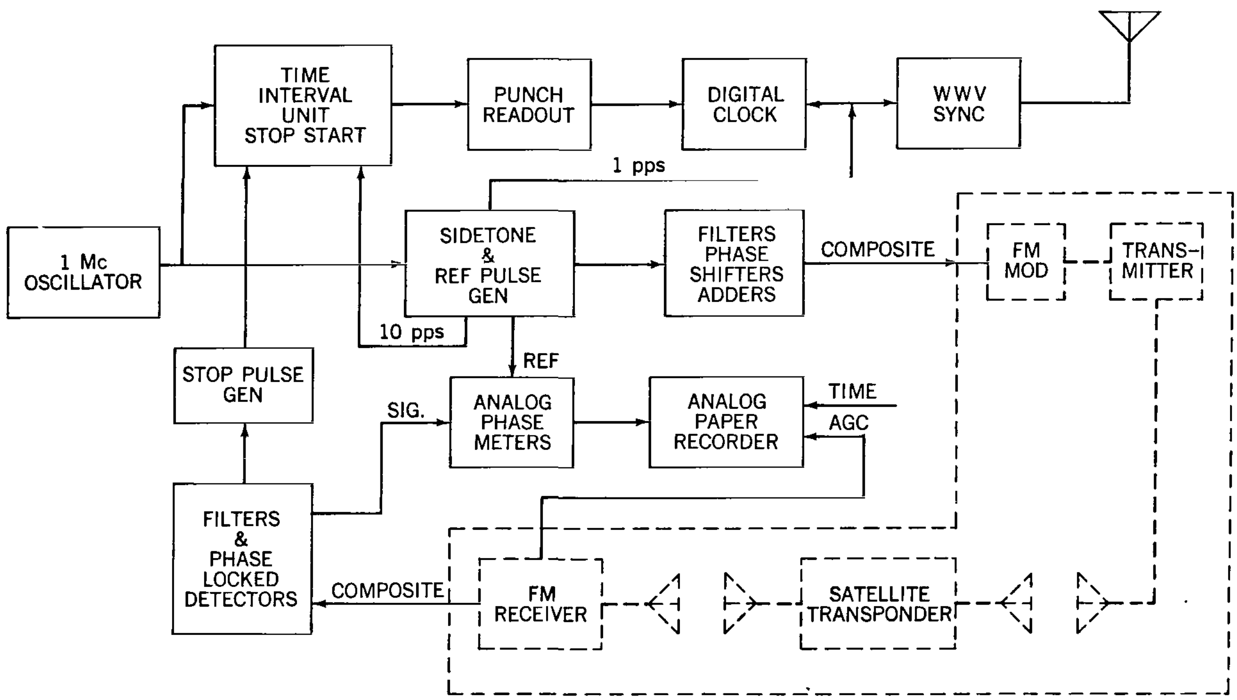
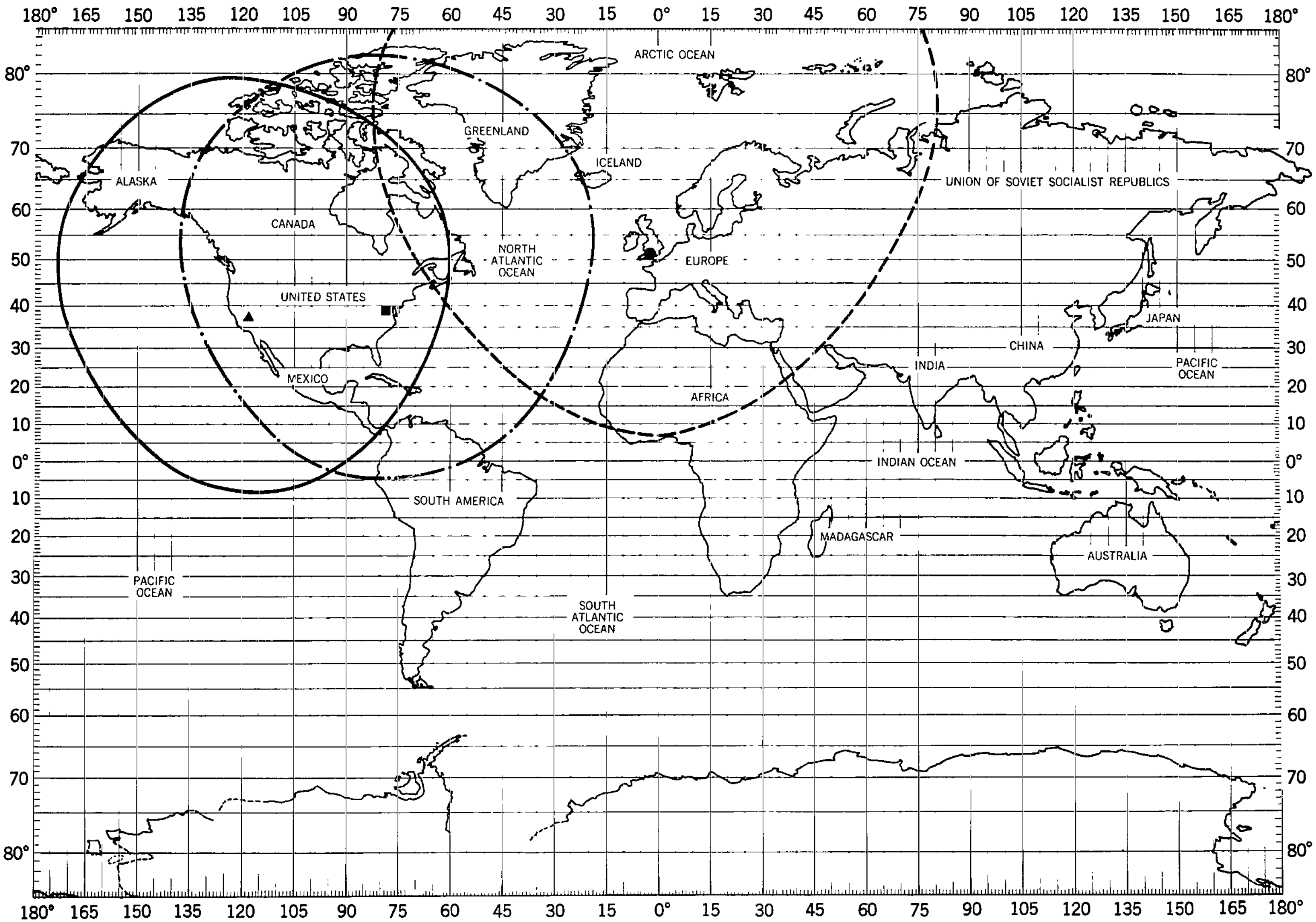


Figure 13—Relay I system block diagram.

Fortunately, the Relay I transponder has two extra channels with narrower bandwidths (2 Mc) than the main experiment channel. These channels were provided for duplex voice communication and can be operated when the main channel is turned off. With these two channels, simultaneous ranging can be performed by two stations. Three systems were originally planned, for use at sites in England, Washington, D. C., area, and Goldstone, California (Figure 14). The calculations for the Sidetone Range Only System for Relay I at 10,000 km are shown in Tables 8 and 9.

Synchronous Communications Satellite (SYNCOM)

SYNCOM is the second of a series of communication satellites in the NASA communications experiment program. Initially it will be a fairly light satellite with moderate communication abilities and will be placed in a 24-hour (synchronous) orbit. Perliminary planning calls for the satellite to be placed in orbit with a 33 degree inclination angle approximately centered on the 30 degree meridian. The transponder will receive on approximately 7 Gc and transmit on 1.8 Gc with a power output



SYMBOL	STATION	CIRCLE OF VISIBILITY	SATELLITES
■	LURAY	— · — · — ·	RELAY I
▲	GOLDSTONE	————	SYNCOM
●	WINKFIELD	– – – – – –	EGO

Figure 14—Radio horizon coverage for Relay I station (1500 mi altitude).

Table 8

Characteristics of the Relay I System for Transmission to Satellite at 10,000 km*

System Parameter	Value
Transmission power, 10 kw	+70 dbm
Satellite antenna gain	0 db
Losses	−3 db
Ground antenna gain 40 ft diameter antenna at 1700 Mc	+44 db
Path loss for 10^4 km at 1700 Mc	−177 db
Received signal power	−66 dbm
Receiver noise figure	11 db
Receiver noise power at 2 Mc bw	−100 dbm
Received signal-to-noise ratio (FM, M.I. = 2)	+34 db

*Synchronous detection of the FM carrier is assumed only for the ranging system.

Table 9

Characteristics of the Relay I System for Transmission to Ground at 10,000 km*

System Parameter	Value
Transmission power, 10 watts	+40 dbm
Satellite antenna gain	0 db
Losses	−3 db
Ground antenna gain 40 ft diameter antenna at 4 Gc	+51 db
Path loss for 10^4 km at 4 Gc	−184 db
Received signal power	−96 dbm
Receiver noise power density at 290°K	−174 dbm
Receiver noise power at 100 cps bw	−154 dbm
Received carrier-to-noise ratio	+58 db
Sidetone signal-to-noise ratio 10 cps bw assuming 10% total power available in sidetone	+58 db

*Synchronous detection of the FM carrier is assumed only for the ranging system.

of 2 watts. Two channels are provided with 500-kc bandwidth. The design of the transponder is so similar to that required for a coherent range rate tracking system that the program manager has agreed to incorporate coherent translation from 7 Gc to 1.8 Gc and also to transmit a multiple of the satellite local oscillator frequency.

The location of the synchronous orbit is such that tracking performed by the Minitrack network will be marginal. Although the accuracies required are not very great—since tracking is mainly needed for station-keeping—range and range rate measurements are still necessary.

It is planned to incorporate the Range and Range Rate System at three SYNCOM ground stations, using the available SYNCOM modulators, transmitters and antennas (Figure 15).

With the two channels available, simultaneous tracking can be performed by two ground stations, time-sharing the satellite with the communications experiment. Thus, accurate angle measurements can be made by triangulation, as well as by the range and range rate data, and complete position information can be obtained.

An example of the calculated system parameters is given in Tables 10 and 11.

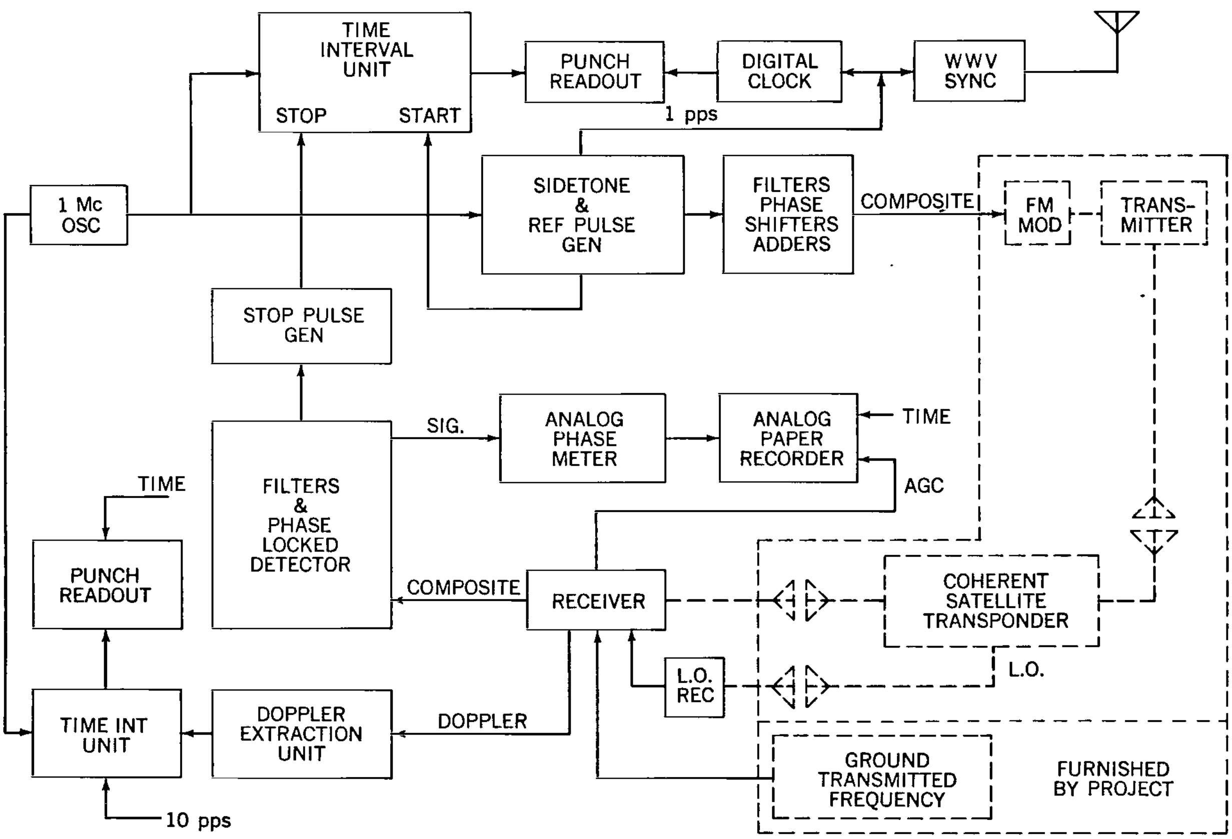


Figure 15—Syncom block diagram.

Table 10

Characteristics of the SYNCOM System for Transmission to Satellite at 40,000 km*

System Parameter	Value
Ground transmission power, 2 kw	+63 dbm
Satellite antenna gain	0 db
Losses	-3 db
Ground antenna gain 30 ft diameter at 7 Gc	+54 db
Path loss for 4×10^4 km at 7 Gc	-202 db
Receiver signal power	-88 dbm
Receiver noise figure	12 db
Receiver noise power at 250 kc bw	-108 dbm
Received signal-to-noise ratio	+20 db

*Synchronous detection of the FM carrier is assumed only for the ranging system.

Table 11

Characteristics of the SYNCOM System for Transmission to Ground at 40,000 km*

System Parameter	Value
Satellite transmission power, 2 watts	+33 dbm
Satellite antenna gain	0 db
Losses	-3 db
Ground antenna gain 30 ft diameter antenna at 1.800 Gc	+42 db
Path loss	-190 db
Received signal power	-118 dbm
Receiver noise power density at 290°K	-174 dbm
Receiver noise power at 100 cps bw	-154 dbm
Received carrier-to-noise ratio	+36 db
Sidetone signal-to-noise ratio at 10 cps bw assuming 10% of total power available in sidetone	+36 db

*Synchronous detection of the FM carrier is assumed only for the ranging system.

Eccentric Orbiting Geophysical Observatory (EGO)

The Eccentric Orbiting Geophysical Observatory (EGO) satellite will be launched into a highly eccentric orbit with an apogee of 100,000 km and an inclination of 33 degrees. This high an apogee is not conducive to tracking by the Minitrack network. It is felt that the Range and Range Rate System can augment the Mintrack network so that accurate position information required by the experimenters can be obtained. Three complete Range and Range Rate stations, and an appropriate transponder are planned for this satellite. The stations will be located to obtain a maximum of data when the satellite is near apogee.

For system sensitivity calculations, refer to the antenna section of this report (page 14).

Agena-B Injection Tracking

It has been suggested that it would be desirable to obtain very precise information on position and velocity during the orbit injection period for the Agena-B vehicle series. Both EGO and the Orbiting Astronomical Observatory (OAO) use an Agena-B, and injection occurs over northern Australia, four Range and Range Rate Systems are proposed with a simplified ground antenna system. The suggested station locations are shown in Figure 16. The system sensitivity calculations are given in Tables 12 and 13 for a 2-foot diameter antenna with a beamwidth of 20 degrees. The greatly simplified antenna system will reduce the cost of the overall system considerably; however, if immediate access to final position information is required, a small, general purpose digital computer will be included at one station.

PRELIMINARY SYSTEM TEST

General

While the concepts for the Range and Range Rate System are still being formulated, a concurrent program for preliminary testing is being carried out by the Goddard Space Flight Center.

A complete Range Only System was designed and constructed to check out the system's capability of automatic ambiguity resolution, to determine its sensitivities for the communication links used, and to verify that the resolution of 15 meters was realizable. The Doppler portion has not been instrumented to date but special receivers for this system testing are expected very shortly. As soon as they are incorporated into a system configuration, that portion will also be tested.

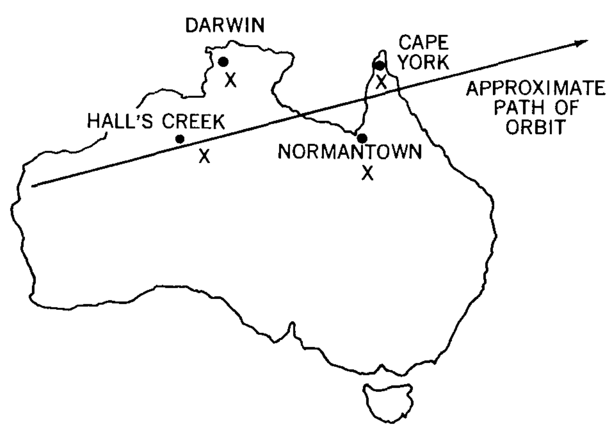


Figure 16—Typical arrangement in Australia for Agena injection tracking.

The configuration chosen is the one at Blossom Point, Maryland (Figure 17). It consisted of a range tone generator consisting of a 1 Mc oscillator stabilized to 5 parts in 10^{10} feeding a system of Packard-Bell transistorized modules; a modified commercial version of an ARC-60 AM transmitter; a Nems-Clarke

Table 12

Characteristics of AGENA Injection for Transmission to Satellite at 1000 km*

System Parameter	Value
Ground transmission power, 1 kw	+60 dbm
Satellite antenna gain	-8 db
Losses	-3 db
Ground antenna gain 2 ft diameter at 1535 Mc	+17 db
Path loss 1000 km at 1535 Mc	-156 db
Received signal level	-90 dbm
Receiver noise figure	11 db
Receiver noise power at 250 kc bw	-109 dbm
Received signal-to-noise ratio	+19 db

*Satellite antenna gain of -8 db is assumed because shroud may still be on satellite.

Table 13

Characteristics of AGENA Injection for Transmission to Ground at 1000 km*

System Parameter	Value
Satellite transmission power, 0.2 watt	+23 dbm
Satellite antenna gain	-8 db
Losses	-3 db
Ground antenna gain 2 ft diameter at 1.7 Gc	+18 db
Path loss 1000 km at 1.7 Gc	-157 db
Received signal level	-127 dbm
Receiver noise power density at 290°K	-174 dbm
Receiver noise power at 100 cps bw	-154 dbm
Received carrier-to-noise ratio	+27 db
Received sidetone signal-to-noise ratio at 10 cps bw assuming 10 percent total power available in sidetone	+27 db

*Satellite antenna gain of -8 db is assumed because shroud may still be on satellite.

AM receiver with a 300 kc bandwidth; a 2-watt phase-modulated solid-state transmitter from Vector Telemetry Corporation; an FM Nems-Clarke receiver modified to produce PM synchronous detection operating at 500 kc bandwidth, an H. P. 524C frequency counter used as time interval unit; an H. P. 561A printer.

The range tone generator produced four sidetone frequencies: 100 kc, 20 kc, 4 kc, and 800 cps. A frequency of 160 cps was contemplated, but was not used for two reasons. First, the transmitter

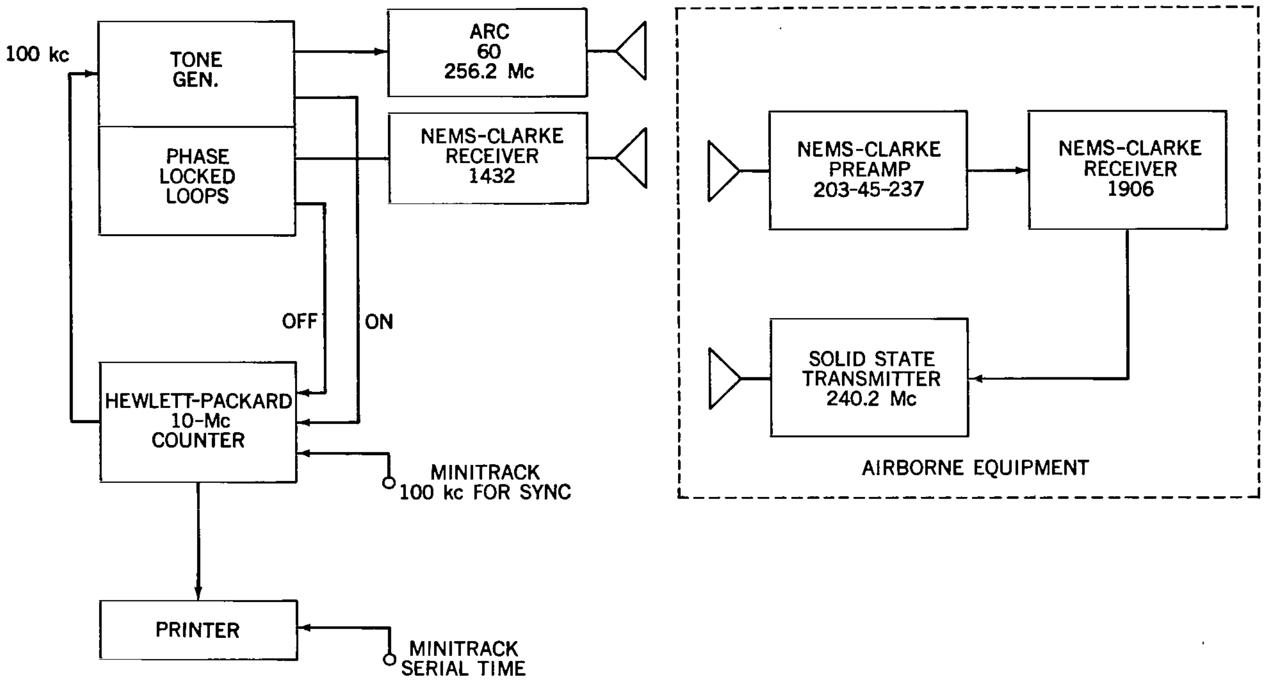


Figure 17—Blossom Point test block diagram.

on hand could not be modulated properly at this frequency, and secondly, the airplane range was not expected to exceed 100 mi so there was no need for the finer (580 mile) ambiguity resolution provided by the 160 cps. The tone generator outputs fed the amplitude modulated transmitter which was transmitting 1 watt at 256.5 Mc. A 10-element TACO yagi array provided 10 db of gain. The signal was received by the airborne equipment in the plane. A preamplifier was used to reduce the noise figure to 3 db in the Nems-Clarke receiver. A phase-modulated, solid-state, transmitter transmitting 2 watts at 240.2 Mc was used to retransmit the received signal back to the ground station. The second Nems-Clarke receiver on the ground was used to receive the retransmitted signal. It was preceded by another yagi, with 10 db antenna gain. This receiver was normally a phase-locked FM receiver with a 500 kc bandwidth, but was modified so that the phase-locked loop was narrowed in frequency to about 200 cycles bandwidth, effectively providing PM synchronous detection.

The modulation indices for the transmitter were adjusted on the ground from the transmission end so that the received signals at the four frequencies were all equal in amplitude when detected by the ground receiver. Each signal was fed through individual active tuned filters and then fed to phase-locked oscillators, as shown in Figure 8.

The remaining system is standard, using a H. P. frequency counter for the 10 Mc time interval unit, a frequency standard, and a H. P. digital printer which printed at a rate of four readings per second. This data was compared to the optical tracking data and results from one run are shown in Figure 18.

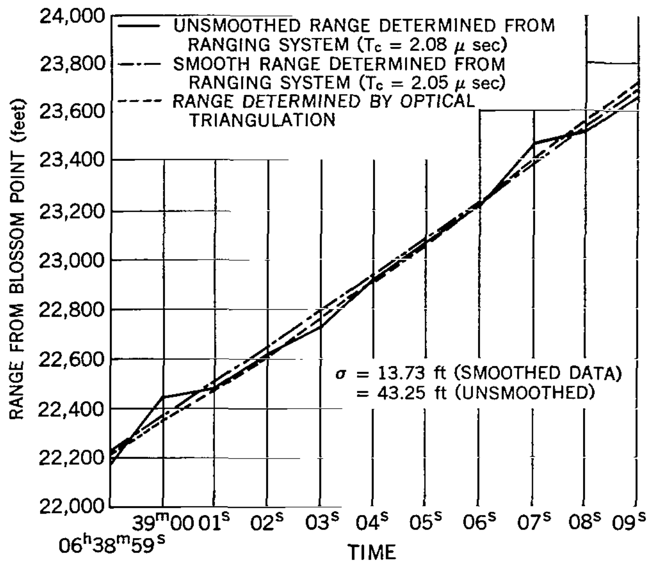


Figure 18—Graph of residuals.

In addition to the automatically ambiguity resolved digital print out, four analog phasemeters were included, consisting of essentially 4 flip-flops and four simple R-C network filters with time constants of 0.1 sec. The phase-lock loops at 100 kc, 20 kc, 4 kc, and 800 cps had corresponding loop bandwidths of 30 cps, 20 cps, 10 cps, and 3 cps. The outputs from the analog phasemeters were recorded on an eight-channel Sanborn recorder along with timing and AGC from the receiver.

Because multivibrators were used for the voltage controlled oscillators at the three lower frequencies, and because the power supply system tended to drift, there were instabilities in the VCO's. Apart from this, the system operated very satisfactorily. Various amounts of attenuation were inserted alternately and simultaneously in both the ground transmitter end and the airborne transmitter end, and again the system behaved as was expected. Spurious outputs from the ambiguity-resolving logic were noted when the received signal levels approached 120 dbm—this would be expected from an AM system operating at such a wide bandwidth. It should be pointed out that the antennas used on board the aircraft were standard commercial UHF stub antennas with probable gains of -5 to -10 db.

Calibration and Data Analysis

To provide a standard by which to calibrate the ranging system, two sidereally-driven astrographic cameras located at Blossom Point and Blossom Point Annex were used to determine the airplane's position optically. The camera at Blossom Point Annex is located approximately six miles due east of the camera at Blossom Point. A second order survey was performed to locate the position of the camera at the Annex with respect to the camera at Blossom Point to an accuracy of within six inches.

The cameras are identical, each consisting of an Air Force aerial reconnaissance lens on a sidereal mount. Each camera has an f5, 40-inch focal length lens, and each covers an angular field of about 11 by 14 degrees. The image is recorded on an 8 × 10 inch glass photographic plate in a holder at the bottom end of the camera tube.

The sidereal mount allows the cameras to rotate about two perpendicular axes which intersect at the nodal point of the lens. The polar axis is aligned parallel to the earth's axis of rotation. The cameras may be rotated about the declination axis and clamped in the desired angular position.

The calibration was performed by flying an airplane on a West-to-East course between the two stations at night (Figure 19). The airplane contained a flashing light (triggered by the Minitrack time code) and the ranging transponder. As the airplane passed over the Blossom Point area the two cameras simultaneously photographed the trail of flashing light images against a star background. These photographic plates were then reduced by the standard techniques developed for Minitrack calibrations. In this manner the direction cosine of the flashing light in the airplane was determined from both cameras, and the range of the light from the Blossom Point transmitter and receiver was accurately determined through triangulation. The airplane was flown at an altitude of 15,000 ft so that the resulting triangle approximated an isosceles triangle. In this configuration the computed range is the least sensitive to small random errors within the direction cosines from the cameras to the flashing light.

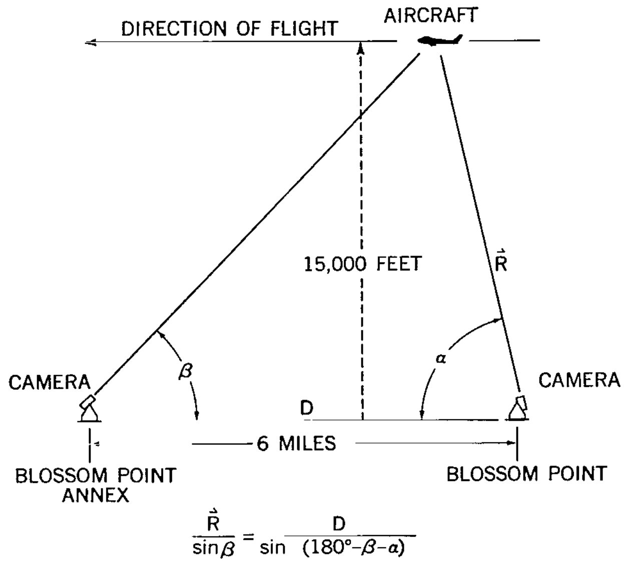


Figure 19—The ground setup at Blossom Point, Maryland.

Ideally the two vectors defined by the set of direction cosines from the two cameras should exactly meet at the flashing light. However, because of small errors in the measurement and reduction of the optical data, that ideal is seldom attained. In such cases the computer solves for the shortest line connecting the two vectors and locates the flashing light at the mid-point of this line. A measure of the accuracy of the optical range determination can be deduced from the length of this connecting line.

The range of the airplane from the Blossom Point ranging transmitter and receiver can be found by the formula

$$r = 49.1679(N-T_c)$$

where r is the range in feet, N is the raw reading of the range counter, and T_c is the electronic time delay within the ranging system measured in units of 10^{-7} sec.

Second and possibly third order time correction terms would have to be included if the system were being used to track a fast moving satellite at a great distance. However, an analysis has shown that this formula is entirely sufficient for the case of an airplane calibration run.

It is the purpose of the calibration to determine the values of T_c such that the parallax corrected range as computed from the optical calculations and the ranging system shall agree as closely as possible. The calibration time constant derived on the basis of a single observation is given by

$$T_c = \frac{49.1679N-R}{49.1679}$$

where R is the computed range determined by optical triangulation.

The T_c adapted as a result of this calibration is the simple average of T_c's determined for each observation.

The H. P. printer used to record the time and counter (N) readings was not functioning perfectly at the time of calibration, and some jitter was observed in the recording of the least significant digit the counter reading. For the reason, as well as the fact that the ranging system has a resolution only to ±50 ft, a second order polynomial was fitted to the raw integral one-second counter readings by the least squares method and the range was determined from the smoothed data.

A calibration time constant of 20.8×10^{-7} sec was computed for the ranging system. Table 14 shows some of the computed results of the ranging calibration. The probable error derived from the smoothed data was found to be 13.73 ft.

Table 14
Tabulation of Ranging Data
($T_c = 20.8 \times 10^{-7}$ sec)

Time	N	$\bar{N}$	r (ft)	R (ft)	r (ft)	ΔR (ft)
$06^h 38^m 59^s$	472	472.650	22,219	22,204	-15	20.0
$39^m 00^s$	477	475.442	22,352	22,342	-10	20.2
01^s	478	478.278	22,494	22,480	-14	1.8
02^s	481	481.158	22,637	22,623	-14	2.3
03^s	483	484.082	22,779	22,769	-10	1.7
04^s	487	487.051	22,927	22,918	-9	2.0
05^s	490	490.064	23,074	23,072	-2	21.6
06^s	493	493.122	23,222	23,225	+3	0.1
07^s	498	496.223	23,374	23,386	+12	1.5
08^s	499	499.369	23,532	23,549	+17	1.5
09^s	502	502.559	23,689	23,716	+27	1.3

N = Uncorrected 10 Mc counter reading.
$\bar{N}$ = Smoothed counter reading (2nd degree least squares pit).
r = Computed range from ranging system using T_c = 20.8.
R = Computed range from optical triangulation.
r = Range residual (R-r).
ΔR = Miss distance in optical triangulation.

ACKNOWLEDGMENTS

The authors wish to thank Dr. F. O. Vanbun for establishing the accuracy needed for such a system, Mr. V. R. Simas for suggesting the method of coherent transponding described in this report, and Dr. Elie Baghdady of ADCOM Inc. for the many consultations provided by and other members of his company.

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Appendix A

High-Accuracy Satellite Tracking Systems

Introduction

The need for greater reliability and accuracy in measuring the range, range rate, direction cosines and angle rates of distant missiles and satellites has recently intensified the need for new radio tracking systems, or for extension of the capabilities of existing ones. New or improved communication techniques are needed to yield these measurements within tight tolerances despite various limiting factors. Several promising approaches are under study by ADCOM, Inc., of Cambridge, Massachusetts, under NASA contract No. NAS5-1187, in close coordination with and in direct support of activities of the Goddard Space Flight Center. The main objectives are to provide information needed for designing distance, velocity, and angle measuring radio systems that meet both immediate and future NASA requirements.

This discussion constitutes the first Quarterly Progress Report and the "Task I Report" specified in the contract, and it presents the results of investigations carried out at ADCOM, Inc., between April 20 and June 30, 1961.

Fundamental Considerations

Radio techniques for measuring ranges, velocities, and relative positions of distant objects make use of three basic properties of wave propagation:

1. Electromagnetic waves travel with a finite and known speed.

2. The instantaneous phase of a traveling wave (with respect to a standard reference) changes linearly with the distance traveled.

3. The frequency f_{rec} received by an observer moving with velocity v *toward* a source which is emitting a note of frequency f_{trans} is given by

$$f_{rec} = \sqrt{\frac{c+v}{c-v}}\, f_{trans}$$

where c is the velocity of wave propagation in the intervening medium.

Property 1 is the basis for radar-type pulsing distance-and-velocity measurement systems. Such systems may or may not use a vehicle-borne transponder, although a system capable of repeating signals to a ground station is generally preferable.

Property 2 can be used in a number of ways to extract tracking data. For example, the direction cosines of a beacon-carrying missile can be determined from the instantaneous phase difference between a signal from the missile as received simultaneously by two antenna pairs equally spaced on short, mutually perpendicular, accurately measured baselines. Similarly, in the sidetone ranging technique, the instantaneous phase shift acquired by tones modulating an RF carrier, after transmission from the ground station, amplification and retransmission (or reflection) at the missile, and final reception back at the ground station can (after correction for spurious phase shifts for the Doppler effect) be used to determine the total distance to and from the missile.

Property 3, the Doppler effect, is the basis for measuring the rate at which a radiating source approaches, or recedes from an observer. If the transmitter frequency is known the preceding equation can be solved for the relative velocity. The Doppler frequency shift $\left(f_{trans} - f_{rec}\right)$ experienced by a narrow-band signal in a receiver that moves with vector velocity $\vec{V}_{missile}$ and receives a wave traveling in vector direction $\vec{n}_T$ from a stationary transmitter is approximated by

$$\frac{\left(\vec{V}_{missile} \cdot \vec{n}_T\right) f_{trans}}{c + \vec{V}_{missile} \cdot \vec{n}_T}$$

for small values of v. If the signal is re-radiated in a new vector direction $\vec{n}_R$ and is re-received by a stationary receiver, the narrow-band signal will experience an additional Doppler shift of

$$\frac{\vec{V}_{missile} \cdot \vec{n}_R}{c + \vec{V}_{missile} \cdot \vec{n}_R} \left(1 - \frac{\vec{V}_{missile} \cdot \vec{n}_T}{c + \vec{V}_{missile} \cdot \vec{n}_T}\right) f_{trans} \; .$$

These elementary properties of traveling waves can be applied in a variety of ways to measure the range, velocity, and direction cosines of a missile. Systems using missile-borne transponders capable of repeating signals to a ground station would seem to provide increased flexibility, simplicity, and reliability in ground reception at the cost of missile-borne power and payload weight. Severe limitations on the weight and volume of missile-borne equipment are imposed by considerations of available thrust and total impulse. An appropriate choice of modulation and signal-processing techniques may greatly increase the reliability and accuracy attainable with a given amount of transponder power.

Spacecraft communications and tracking are handicapped also by the uncertainties of propagation. Foreknowledge of the nature of the perturbations imposed on the signal in transit aids the choice of modulating waveforms most easily distinguished from, and most immune to, fluctuations in transmission. Even in the absence of sufficient data on this subject, however, it is possible to anticipate the likely difficulties and select modulation waveforms that can be isolated from disturbances with a minimum of error in identifying the information-bearing characteristics.

There are also important uncertainties in our knowledge of the speed of light: the most refined measured value known today fixes it at $c = 299792.5 \pm 0.4$ km/sec. Since the speed of light enters vitally into all the basic propagation computations, the inaccuracy in the value of c ultimately limits the accuracy of all radio tracking measurements. For example, in the measurement of range, if we ignore the contributions of all other potential sources of error, the range error ΔR caused by the uncertainty ± 0.4 km/sec in c is approximately $1.33 \times 10^{-6} R$. If $R = 250{,}000$ mi, then $\Delta R \approx \pm 1/3$ mi.

Finally, there is the inevitable problem of circuit stability and reliability, especially where reference oscillators form an essential part of the system.

Survey of Radio Tracking Systems

Angle Measurement

The basic method of angle tracking requires the measurement of the phase difference between the carrier signals as received simultaneously at separate antennas from a single remote source. The carrier frequency and antenna separation (baseline length) are design parameters; the measured variable is the carrier phase difference. Provided that the baseline is very short compared to the distance to the object being tracked, (so that the arriving rays are essentially parallel) the cosine of the angle included between the line of sight and the baseline may be calculated as follows:

$$\cos\theta = \phi_d c/\omega_c D \ ,$$

where ϕ_d = phase difference between received signals (radians), $c = 299792.5 \pm 0.4$ km/sec., ω_c = carrier frequency (radians/sec), D = baseline length (km). There are practical limits to the precision of direct measurement of small phase shifts. Long baselines and high carrier frequencies are therefore indicated, in order to provide increased differential carrier phase shift.

The Minitrack system measures the ratio ϕ_d/ω_c in the equation above. This ratio represents the time delay between the reception of corresponding phases of the carrier wave at the separate antennas Because of practical limits on the resolution of small time delays, use is made of the heterodyne principle, which provides that relative phase shifts may be preserved while translating signals in frequency. The Minitrack system translates the incoming 108 Mc signals to 500 cps signals for phase comparison. Because of the preservation of phase shifts, the differential time delay of reception is magnified by the ratio $108 \times 10^6/500 = 2.16 \times 10^5$ prior to the time delay measurement. The digital phasemeter counts the pulses from a standard 500 kc source during the delay interval measured; hence, it has a delay time resolution of 2 μ sec. Referring this resolution limit to the 108-Mc signal reduces the differential delay resolution at the antennas to $2 \times 10^{-6}/2.16 \times 10^5 = 9.25 \times 10^{-12}$ sec. Differential time delay resolution may be translated directly into differential range resolution as follows:

$$\Delta R = c \Delta t = 9.83 \times 10^{8} \times 9.25 \times 10^{-12}$$
$$= 9.1 \times 10^{-3} \text{ foot.}$$

For the 500-foot baseline employed, the resolution of direction cosines becomes $9.1 \times 10^{-3}/500 = 1.82 \times 10^{-5}$. For angles near the zenith, the resolution then becomes 3.7 seconds of arc.

The AZUSA tracking system similarly employs the heterodyne principle to preserve differential carrier phase shift at a frequency of 2 kc for fine angle measurement. In addition, the phase shift is magnified by frequency multiplication of the 2-kc signal to 320 kc. The phase shift at 320 kc thus becomes 160 times as great as the original carrier phase shift. It is necessary to resolve the phase shift at 320 kc only to the nearest 1/2 cycle. This is equivalent to resolving the 5-Gc carrier phase shift to 1/320 cycle. In terms of range difference, resolution is thereby provided to about 0.0003 ft.

A baseline of about 164 feet is employed by AZUSA for the fine angle measurement. Hence, the fine direction cosine resolution becomes $0.0003/164 \pm 2 \times 10^{-6}$.

The MISTRAM system uses a 10,000-foot baseline together with an 8-Gc carrier to provide a range difference accuracy of 0.03 ft. The technique employed for phase measurement has not been disclosed in available documents.

To obtain accuracies of angle measurement commensurate with the available resolution, it is necessary to maintain stable carrier frequencies and baseline lengths. Short baselines may be stabilized through choice of materials of construction for waveguides, temperature control, and protection of guides from mechanical stress. MISTRAM, because of its long baselines, employs in addition a correction technique whereby the two-way phase shift of waveguides is continuously measured and used to correct for changes in one-way phase shifts.

Range Measurement

The basic principle employed for range measurement is that the two-way phase shift of an RF carrier or a modulation of the carrier is proportional to the range. AZUSA and MISTRAM measure carrier phase shifts for high-resolution range and range rate determination, and modulation phase shifts for ambiguity resolution. AZUSA employs three harmonically related waves to modulate the carrier frequency: 98.3 kc, 3.93 kc, and 157 cps. The measured phase shift of the received 98.3-kc component can provide unambiguous range data, except during temporary signal loss. The two lower frequencies are used only as needed to resolve coarse and intermediate range ambiguities. A ranging system under development by the Goddard Space Flight Center proposes the use of 5 or 6 harmonically related waves (500 kc, 100 kc, 20 kc, 4 kc, 800 cps, 160 cps) to provide a single-sideband modulation of the carrier. Here, again, the measured phase shift of the highest component, 500 kc or 100 kc, provides unambiguous range information, with the provision that the lower-frequency terms may be employed to resolve ambiguities resulting from signal loss or lack of initial data.

The MISTRAM system for unambiguous ranging uses a ramp function to sweep the carrier from 7.884 Gc to 7.892 Gc. During the sweep modulation, the cycles of phase difference between the transmitted and received carriers are counted. Since the maximum frequency deviation is known to be precisely 8 Mc, the average range during the sweep may be determined from $R = Nc/(4\pi \times 8 \times 10^6)$, where N is the number of cycles counted and c is the velocity of propagation. Note here that if only *whole* cycles are counted, the resolution of range ambiguity is limited to 9.77 ft. To provide for the published system range accuracy of 0.4 ft, some means not presently apparent must be employed to refine the measurements of carrier phase shift.

The Minitrack system provides no ranging data, since no communication exists from the ground to the satellite.

The importance of preserving carrier phase information in the transponder is recognized in both the AZUSA and MISTRAM systems, where phase-locked loops provide for retransmission of coherent carriers having constant frequency offsets from incoming signals. In addition, the AZUSA transponder plays a part in an overall system automatic-frequency-and-phase-control loop. By this system, the ground-received carrier is maintained at 5 Gc, whereas the ground-transmitted carrier contains the Doppler information. An apparent advantage is that angle measurements may be referred to a constant carrier wavelength.

The transponder proposed for the Goddard Space Flight Center single-sideband ranging system must handle simultaneously the signals from three ground stations. It is necessary to preserve the phase relationships between each carrier and its associated ranging sidetones. However, it is not necessary to maintain carrier phase information, since range measurements will be made only with the modulation frequencies.

The Signaling Problem in Sidetone Ranging

Having briefly surveyed the fundamental physical principles employed in radio tracking, and the approach taken in well-known systems to the application of the fundamental principles, we now consider the technique of *sidetone ranging* and examine the signal design problem involved. The ultimate signal that will be transmitted to and from the satellite is here considered to have an RF spectrum derived in some reversible manner from the sum of a number of discrete tones. In application to radio tracking systems, the signal design problem may therefore be concerned with:

1. the choice of the individual frequencies, the relationships among the frequencies, the initial relative phases, and the amplitude weightings of the baseband tones; and
2. the modulation technique to be employed in generating the radio-frequency spectrum from the sum of the baseband tones.

We shall begin with an examination of the modulation problem first, assuming that a specific choice of sidetones has been made, and lead up to the considerations that should govern the choice of the sidetones. For convenience we might start by posing the following question: "If the baseband time function is made up of K discrete tones of equal amplitudes, and if the desired information is carried in the phases of the tones, what is the best modulation technique to employ in transmitting these tones to and from the satellite?"

The important considerations in evaluating the alternative modulation methods are:

1. efficiency of utilization of available (transponder) power in conveying the tones,
2. simplicity of system implementation,
3. signal bandwidth requirements,
4. peak factor of the signal, and
5. the effect of the associated demodulation process upon the spectral distribution of the noise power.

Let us justify briefly each of these points:

The question of the efficiency utilization of available power arises from the fact that a limited amount of power is available in the transponder for investment in the structure of the signal.

The emphasis on simplicity of system implementation is motivated by the desire to minimize the amount of signal processing necessary to extract the desired information, and hence to minimize the potential sources of error and distortion in the circuitry.

The interest in the signal bandwidth requirements is based upon the desirability of minimizing the noise bandwidth, especially in the transponder, and of avoiding dependence on wide-band transmission filters with their relatively greater possibilities of drifting phase characteristics.

The peak factor (i.e., the ratio of the peak value to the rms value) of the signal is important because the ultimate performance of a communication system is usually determined by the average signal power, while the peak power is often a limiting factor in transmitter design.

As for the effect of the demodulation process upon the power spectral density of the ambient noise, it is important to remember that the demodulation process in some systems (e.g., FM) will alter the noise power spectral density in such a way as to favor certain frequencies over others.

Linear Modulation Systems

Amplitude modulation makes inefficient use of available power because a major fraction of the signal power resides in the carrier, and half of the remainder resides in a redundant sideband. This inefficiency secures advantages that are either not exploited (e.g., simplicity of the detector for high signal to noise ratios) or are not sufficiently important in radio tracking systems. The value of these advantages actually vanishes under severe noise conditions, and AM becomes especially undesirable. If the desired sidetones are extracted directly from the IF spectrum by product demodulation and phase-locked oscillator means, then a considerable amount of potential signal power will not be utilized. Other linear-modulation systems are superior to AM in this regard.

Therefore, since it is either comparable with DSB and SSB or is inferior to them, AM is ruled out.

In general, double-sideband (DSB) suppressed-carrier operation may have an advantage over single-sideband (SSB) operation from the standpoint of the peak factor of the signal with peak-power-limited transmitters (as well as a potential 3 db advantage against random noise resulting from the redundant sideband). But this possible advantage may be negated with appropriate initial phasing of the baseband tones. On the other hand, SSB operation uses the minimum permissible bandwidth, avoiding the detrimental effects (likely in DSB) of unsymmetrical phase shifts of corresponding upper- and lower-sideband components by concentrating the available signal power in one sideband.

With regard to the other considerations, SSB is either equivalent to DSB or superior to it.

With or without a residual carrier component, the SSB system is the most commendable among the linear modulation systems. However, a potentially serious disadvantage of SSB, as opposed to the exponential modulation systems discussed next, lies in the limitation on the system's noise performance imposed by an undesirable signal peak factor in a peak-power-limited transmitter. For a signal described by

$$s(t) = \sum_k A_k \cos\left[(\omega_c + k\omega_m)t + \phi_k\right]$$

the peak factor is given by

$$PF = \sqrt{\frac{\left[A_0 + \sum\limits_{k=0} A_k \cos(\phi_k - \phi_0)\right]^2 + \left[\sum\limits_k A_k \sin(\phi_k - \phi_0)\right]^2}{\sum\limits_k \frac{A_k^2}{2}}}$$

where the initial phases ϕ_k of the components have been chosen so that the envelope of $s(t)$ attains an absolute maximum at $t = 0$. It is clear that the peak factor can be reduced by an appropriate adjustment of the relative phases $(\phi_k - \phi_0)$ and/or the amplitude factors A_k.* But if the initial relative phases are adjusted for a minimum peak factor for the signal radiated to the transponder, the phases of the sidetones extracted from the transponder-reradiated signal should be *readjusted* to correspond to the phase relationships that ensure the greatest simplicity and reliability of the final ranging-information extraction and calibration operations. This readjustment is greatly facilitated by the fact that ground transmitter and receiver are usually located next to each other.

Exponential-Modulation Systems

Viewed as a sinusoid that moves back and forth on the frequency scale (in accordance with the instantaneous value of the resultant of the baseband tones for FM, or its derivative for PM), an exponential-modulation signal can be readily seen as an effective carrier of tracking information because its detectability in the presence of random-fluctuation noise is ultimately limited by the noise power spectral density, not the total noise. Although wasteful of frequency space when compared with SSB, exponential modulation offers the following advantages over SSB:

*Contrary to initial suspicions, a choice of $\phi_k - \phi_0$ and of A_k that corresponds to what would obtain in the spectrum of a baseband square wave is very disadvantageous for SSB operation, especially when compared with DSB operation.

1. The satellite repeater design is simplified because the signal can be amplitude limited; hence, no peak-factor problem.

2. Class C radio-frequency power amplification can be used in the transponder as well as on the ground; this improves the efficiency of available power use.

3. The exponential modulation signal offers a degree of flexibility in the delivery of tracking information which is not available in SSB. This will be explained presently.

4. An interesting by-product of the flexibility just cited is that the desired FM tracking signal can be generated more easily than the corresponding SSB signal.

In the present application, the desired baseband sidetones can be extracted from the exponent-modulation signal in at least two ways:

1. by some exponential-demodulation means followed by appropriate phase-locked oscillator isolation of the desired tones; or

2. by direct operation on the IF signal spectrum in a manner similar to that used with SSB.

With high signal-to-noise ratios, the use of exponential-demodulation techniques prior to isolating the desired tones can be very advantageous. If a phase demodulator is involved in the operation, the baseband noise spectrum varies with frequency exactly in the same manner as the IF noise spectrum varies with frequency deviation from the instantaneous signal frequency. For example, white noise plus a carrier at the demodulator input leads to white noise at the demodulator output. But with a frequency demodulator the output noise power spectral density is given by the IF noise spectral density as a function of the frequency deviation from the signal's instantaneous frequency, weighted by the square of this frequency deviation. Thus, white noise plus a carrier at the FM demodulator input leads to baseband noise having a parabolically rising power spectral density with increasing frequency.

These output noise characteristics of FM and PM demodulators in the case of high input signal-to-noise ratios must be seriously considered in selecting the amplitudes of the baseband sidetones before the modulation process. First, since the higher-frequency sidetones are used for deriving more detailed range information than are the ambiguity-resolving components, it is logical to favor these tones, whose higher fraction of the signal power will better enable them to override the noise. Thus, even in the case of SSB and PM output white noise, it is advantageous to invest more power in the higher-frequency sidetones than in the ambiguity-resolving sidetones. The desirability of this type of intensity distribution among the sidetones is even more strikingly obvious in the case of FM demodulation.

The baseband sidetone amplitudes should, therefore, be chosen *not equal.* Rather, these amplitudes should rise with frequency in accord with (1) the accuracy required in identification of the sidetone's phase, and (2) the expected post-demodulation ambient noise spectral density in the vicinity of the desired sidetone frequency. However, when operation in the vicinity of the demodulator's noise threshold is expected, the foregoing recommendation should be reconsidered in a new light: Increasing the amplitudes of the higher-frequency sidetones widens the bandwidth occupancy of the FM or PM signal. This in turn calls for wider IF filter bandwidths, and hence increased noise threshold.

An interesting way to avoid the "parabolic" spectral density of the FM post-demodulation noise in the case of high signal-to-noise ratios would be to frequency modulate at the transmitter with a set of sidetones whose initial phases are set 180 degrees away from the desired reference phases, phase demodulate the FM signal back at the ground receiver, and then use phase-locked oscillators to isolate the desired sidetones. The phase-locked oscillations with a zero static phase error would bear a 90 degree relative phase relationship with the tones at the phase demodulator output, and hence should be immediately suitable (without further phase changes) for phase-comparison operations leading to extraction of the desired ranging information.

The desired ranging tones can also be extracted from the *spectrum* of the exponent-modulated signal directly and without using an exponential-demodulator, in exactly the same manner as in SSB. This manner of baseband-tone extraction may be desirable for several reasons:

1. When the signal-to-noise ratio is well below the demodulator threshold, the action of an FM demodulator (for example) upon the sum of signal and noise leads to a substantial suppression of the signal and of the signal modulation. Consequently, even though the desired tones may be present in the demodulator output deep in the noise, they can be expected to be less favorable to work with than the spectral components in the IF signal.

2. An exponential-modulation signal that is suitable for sidetone ranging can be generated from one basic tone (or, if desired, two harmonically related tones) which frequency- or phase-modulates the desired carrier. Such an approach both permits amplitude limiting and class C power amplification in the transponder, and significantly simplifies the signal-generation problem in the ground transmitter. As will be shown shortly, such a signal can be designed to have a set of spectral components whose initial relative phases and amplitude distribution make them particularly suitable for use in ranging measurements. Moreover, the spectrum will be rich in components spaced sufficiently far from the carrier frequency to be suitable for more and more accurate ranging when the signal-to-noise ratio is high (as is fortunately the case for ranging below 3000 mi or so).

But the proper extraction of desired ranging tones from a signal of the type described here must be made directly from the IF spectrum.

The price for the advantages of this approach is of course bandwidth and inefficient utilization of available signal power. A 3-db improvement in performance against random noise may be achieved by using both the upper and the lower sideband components as in DSB. Care must be taken not to choose a modulating frequency (or frequencies) that results in an overcrowded spectrum for the resulting signal and thus complicates the problem of isolating the desired tones.

In order to substantiate some of the foregoing statements about the use of a single-tone or a double-tone exponential-modulation signal, let us review briefly the spectral properties of such a signal. First, recall that for the single-tone case,

$$e_{exp}(t) = A_c \cos\left[\omega_c t + \phi_c + \delta \sin\left(\omega_m t + \phi_m\right)\right]$$

$$= A_c \sum_{n=-\infty}^{\infty} J_n(\delta) \cos\left[\left(\omega_c + n\omega_m\right) t + \phi_c + n\phi_m\right] .$$

The important properties of this spectrum are

(1) $J_{-n}(\delta) = (-1)^n J_n(\delta)$

(2) The initial phase $\phi_c + n\phi_m$ offers a potentially useful control over the relative phasing of the components.

(3) $J_n(\delta)$ as a function of n undulates (assuming positive as well as negative values) with gradually increasing amplitude all the way from $n = 0$ to the last maximum (which is absolute) just to the left of $n = \delta$, and then decays quickly toward zero.

Additional flexibility in the properties of the spectrum can be gained by using two tones instead of only one, but a detailed treatment of this matter will not be undertaken here.

Pseudo-Random Waveforms

The use of pseudo-random waveforms or binary sequences in signaling for communication and/or tracking purposes is of particular interest as a potentially effective technique whenever one or a combination of the following goals is a requirement:

1. Maximum immunity to deliberate or accidental jamming.
2. Maximum security from interception by an agency seeking to determine, for example, the most effective counter-measures against the system.
3. Combatting some special propagation (multiplicative) and/or additive noise disturbances.

The price for resorting to pseudo-random signaling techniques can be quite heavy with regard to equipment complexity, but this is not necessarily true of the transponder. The fact that the ultimate ground transmitter and receiver are usually side-by-side (in the applications contemplated in the present program) allows substantial simplifications in instrumenting such a system. But no real and compelling justification is presently apparent for resorting to pseudo-random signaling as opposed to "harmonic" signaling of the type under discussion. However, psuedo-random signaling will not be excluded from further consideration in the present program, especially as more information is gathered on the anomalies of propagation.

Total Versus Incremental Tracking

In the preceding sections we have examined methods for implementing a relatively fixed scheme of range determination. In the present section we shall look at the entire measurement problem from a broader viewpoint. In general, it is possible to distinguish between two basic tracking techniques: In one technique, which we might call "total tracking", the measuring system provides complete tracking data (up to some maximum range) *at each reading*, making use of no previously gathered information. Such a system could follow an object which "jump around" (makes large, random changes in range between measurements). In the other tracking technique, which we might call "incremental tracking," the measuring system capitalized upon the fact that the tracked object can be expected to follow a smooth trajectory and therefore *only the changes in that trajectory* need be measured. Such a system might start from previously acquired data and thereafter it accumulates the incrementally gathered data. A "mixed" system would make occasional ambiguity resolving measurements to maintain accurate reference for the tracking data.

Evidently, when objects move smoothly an "incremental tracking" system should be more efficient, as the following computation indicates. Suppose that a ranging system is to measure and determine, 10 times per second, within which 10-foot interval of range an object lies, for ranges from zero to 30,000 miles. Suppose also that the rate of change of the object's range will never exceed 10 miles per second. A "total-tracking" system (which cannot use velocity information) will require an information rate of about 240 bits per second. (There are about 2^{24} 10-foot intervals in 30,000 miles, and this 24-bit information is sent ten times per second). In a comparable "incremental tracking" system, we would send tracking data 10 times per second, giving information only about changes in position. If in addition, to be conservative, we send complete range information once per second, the information rate is now only about 114 bits per second. (We send 24 bits of range data once per second and realize that in 0.1 second the position can change by at most ±1 mile or a total of about 2^9 10-foot intervals, so that each "tracking" measurement requires only 10 bits.) If constraints on acceleration can be efficiently incorporated into the "incremental" system, a much greater saving in channel capacity can be achieved.

A "total tracking" system, though less efficient, is likely to be simpler to instrument than an "incremental" system. Continuous sidetones would seem to perform as well as any other type of signal in such a system. To increase efficiency, the coarse ranging (lower-frequency) sidetones should be reduced in amplitude and performance regained by averaging over longer periods the phase shifts of these tones. However, efficiency will be lower than for an "incremental" system, and if the power in the coarse sidetones is made too small, system reliability will suffer.

An example of a particularly interesting "mixed" system is the use of switched-sidetone ranging, also called "frequency hopping," with coarse-ranging sidetones switched in for ambiguity resolution only as the occasion demands. This system appears to have the greatest efficiency among the systems considered here and, for very deep space probe applications, represents the most effective way to use the available transponder power. Ideally, only one tone would be sent at a time, and reference would be obtained from a bank of local phase-locked oscillators in the ground receiver whose phase behaviors extrapolate (on the basis of behavior noted while a sidetone is on) the slowly changing phases of corresponding tones while a tone is cut off. Corresponding sharp filters in the transponder (ideally, phase-locked loops whose outputs are interrupted in the absence of input tones) minimize the amount of noise fed to the power amplifier.

Additional representative examples of "mixed" systems are AZUSA and MISTRAM, both of which derive incremental data from measurements of the carrier phase shift. The AZUSA system employs FM sidetones for determination of "total" ranging data, when required. "Total" ranging by the MISTRAM system is accomplished by periodically offsetting the carrier frequency.

Transponder Design

The major (and interrelated) criteria for transponder design and operation have been assumed to be simplicity and reliability. We must also always bear in mind the existence of power and weight limitations. These criteria and limitations suggest that the transponder be little more than a simple, wide-band linear repeater. One additional feature, automatic gain control, probably should be incorporated so that the output power capabilities of the transponder are fully used regardless of input signal level. While it might also seem desirable to conserve power by shutting off the transmitter when no useful input is detected by the receiver, the advantage of such a scheme must be weighed against a possible decrease in system reliability and increase in transponder weight and power consumption.

Now let us consider the problem of noise received in addition to signal at the transponder input. The amount of this noise which is amplified along with the signal and appears in the transponder output will evidently depend on the bandwidth of the circuitry. In particular, it is reasonable to assume that for linear circuitry, the noise power (with no limiting or no automatic gain control) appearing in the transponder output is directly proportional to the bandwidth. For example, suppose that a simple wide-band transponder has a noise bandwidth of 1/2 Mc. The reduction in noise power (no agc) that could be achieved by narrowing the effective noise bandwidth to 5 cps (say, by using phase-locked oscillators as tracking filters) would then amount to 50 db. This improvement would scarcely be noticeable for high signal-to-noise ratio transponder inputs, but at low signal-to-noise ratios and with agc much of the 50 db would appear as an increase in useful signal power (contrasted to noise) at the transponder output.

Even assuming that such an improvement as that described above would be very useful, it is necessary to consider the effect of more complicated signal processing upon system reliability. In the above example, a faulty phase-locked oscillator could make the system completely unuseable. To show how system reliability might be at least partially restored, the transponder could have two modes of operation. One, the "normal" mode (in which the transponder always operates, in the absence of other information) is a simple, side-band repeater mode. To this circuitry may also be attached tracking filters (e.g., phase-locked oscillators) which normally sweep a localized frequency band, searching for a signal carrier and/or sidetones. If these filters find a signal and lock onto it, decision circuits could determine that the filters were both operating and not sweeping, and the transponder could then be switched to a narrow-band mode. Any failure or loss of lock would cause a return to the wide-band mode, so that the system should not break down completely because of failure in the additional signal-processing circuitry.

Another possible modification of simple transponder design is the introduction of nonlinearities into the signal-processing circuitry. The particular example we wish to discuss is the employment of class C amplifiers in the transponder transmitter. The power efficiency of class C stages does exceed that of linear stages by a relatively small amount; however, such an efficiency gain may be more than outweighed by a degradation in system performance, especially with linear-modulation (e.g., SSB) signals, but the same is not true for exponential-modulation signals, as we shall now explain.

Suppose that some type of exponential modulation (say, FM) signaling is employed. Then class C amplification provides limiting action and a certain degree of automatic gain control. The limiting action can suppress noise relative to signal if the received signal-to-noise ratio is high; it is also very possible for the noise to suppress the signal (but to a negligible extent) at very low received signal-to-noise ratios. Hence, in FM signaling systems, class C amplification is more desirable from the viewpoint of transmitter power efficiency than it is objectionable from the viewpoint of noise.

In the case of linear-modulation signaling systems (e.g., SSB), however, the situation is clearly unfavorable to class C amplification. At low noise levels, the resultant limiting action can produce spurious sidetone components that may interfere with phase-comparison measurements to a very objectionable extent. To a less consequential extent, it is also probable that limiting will alter the relative amplitudes of carrier and sidetone signal components. At high noise levels there exists the additional possibility of relative signal suppression by noise, much as in the preceding discussion of FM.

In view of the preceding remarks, we recommend that class C amplification be employed in the transponder in exponential-modulation (FM, PM) operation, but *not* in SSB operation.

Implementation of Sidetone Ranging

The range information to be derived from a SSB sidetone ranging signal is contained in the differential phase shifts of the sideband components relative to the carrier phase shift. Three problems are immediately apparent with regard to reliable recovery of range data: (1) obtaining precision in the measure of phase shifts, (2) resolving ambiguities, and (3) reducing the uncertainties of extraneous phase shifts.

A straightforward approach to the attainment of precision is through the use of digital techniques. It is proposed, therefore, to express phase shift of the demodulated sidetones as time delays and to make the smallest digital increment of time as short as possible. Representative of such a subsystem for measurement of time intervals is the HP5275A counter, operating in conjunction with an HP101A frequency standard. The minimum time increment that may be resolved by typical equipment is 10 nanoseconds. The equivalent resolution of range is 4.92 ft, independent of the frequencies of the sidetones employed.

Selection of the number of sidetones to be employed and the frequencies of these tones should be made with regard to accuracy of ranging, including the resolution of ambiguities. The choice of 160 cps for the lowest frequency term will provide for unambiguous ranging up to 508 n. mi. If it were presently possible to measure the phase shift of 160 cps with high accuracy, this single tone would suffice for our ranging requirements. However, the following example will indicate the need for other ranging tones as well. Let the system ranging accuracy be specified as ± 10 ft. Because of the uncertainty of 4.92 ft. allotted to the time measurement, there would remain a maximum allowable uncertainty of about 5 ft ascribed to the measurement of 160-cps phase shift. Referred to the digital time delay measurement above, the tolerance becomes about ± 10 nanoseconds. The allowable phase error then becomes

$$\begin{aligned} \Delta\phi &= \pm 360 \times 160 \times 10^{-8} \text{ degrees} \\ &= \pm 5.76 \times 10^{-4} \text{ degrees.} \end{aligned}$$

It is not expected that drifts in the parameters of circuits which handle the 160-cps information could be maintained within these limits. The allowable phase error increases directly with the sidetone frequency. For example, a 500-kc tone could have a maximum of ±1.8 degrees of phase error. Provided that circuit developments can insure such performance, the specified ranging precision is attainable.

Accuracy requirements imply also that ambiguous data must be eliminated. A technique by which this may be accomplished requires that the phase shift tolerance on each of the lower frequency sidetones be sufficient for identifying a whole cycle of the next higher tone employed. The proposed system would employ successive tones with 5:1 frequency ratios. With a tolerance of about ±1.8 degrees on the phase error of the 500-kc tone, the phase tolerance for 100 kc would be slightly less than 36 degrees, and for the other lower tones would be about 30 degrees.

The process of extracting the demodulated sidetones for obtaining ranging precision and resolution of ambiguities requires tone filter. Phase-locked oscillators have been suggested for their properties of narrowband filtering combined with frequency tracking. It appears that such filters as presently conceived may function adequately for extraction of the lower sidetones. However, considerable improvements in stability and in reduction of static phase errors are needed in order to enable such filters to function with the highest frequency tone.

It may prove more satisfactory to avoid or simplify, rather than to solve directly, some of the problems caused by the need to measure small phase shifts between sinusoidal signals. In particular, if errors in phase-shift measurement are primarily due to random noise, it will generally be advantageous to increase the range-signal frequency up to a point where there need only be about three distinguishable phases per cycle, instead of the 200 required for a 500-kc highest-frequency sidetone. An argument supporting this statement follows:

Suppose that the range-signal frequency is multiplied by n. Then the number of distinguishable phases per cycle of shift need only be $1/n$ as great as before to attain the same range precision. Assuming that the signal power is unchanged, we are then effectively multiplying lengths of the phase message by n, or increasing the phase-message power by n^2. Other things being equal, the system bandwidth, hence also the noise power, must increase by n. But this means that the effective message-to-noise power ratio of the system is increased by n, which would either increase system reliability or permit a reduction in transmitter power by n to achieve the same reliability. Thus the ranging-signal frequency should be made as high as possible, at least to the point where the number of distinguishable phases per cycle is quite small, perhaps as small as three. With only two it would be impossible to tell, without Doppler measurement, whether relative phase was increasing or decreasing. Our major assumption in the foregoing argument is that the system is not marginal, that is, that the noise peaks seldom if ever are large enough to cause the phase to jump a cycle.

Angle Measurement

Although major effort is being directed to the development of a precise ranging system, it is of interest to investigate the measurement of angular position for several reasons. Ultimately the angular position of the satellite being tracked will be derived from the range information collected by a number of ranging stations. It would be instructive to investigate the measurement of angular position at a single station if only for comparison with the method presently under consideration at Goddard Space Flight Center. However, there is another consideration which enters the problem of position measurement. If some form of repeating satellite is to be interrogated by a ground station, it may be necessary to use a highly directional transmitting antenna which must be given up-to-date information on the angular position of the satellite in order to allow it to be directed toward the vehicle. This information can be derived in three ways:

1. A prediction can be made of the satellite's orbit and the antenna directing mechanism can then be programmed to track the satellite; or
2. Angular position can be computed in real time at a data collecting station which uses the range information from the system of ranging stations and feeds back to these stations angular-position information; or
3. Each station can generate its own angular-position data by using a local short-baseline angle-measuring system.

The method of tracking with a predicted program is undesirable from the standpoint of accuracy, and it requires considerable intital information. It could not be used reliably for tracking objects while they are being placed in orbit because of uncertainties in launching.

The second alternative—remote real-time data processing—requires reliable long-range communication links. The additional system complexity is located in the data-reduction equipment, and all stations are dependent on the control center and are partially slaved to it.

The third method allows each station to operate independently but makes the individual stations more complex. Since it appears that real-time angular measurements must be made to facilitate tracking, it is necessary to compare short baseline and ultra-long baseline systems.

The ranging system under consideration by Goddard Space Flight Center at present is an ultra-long baseline system in which pairs of ranging stations define baselines that are not necessarily perpendicular or coplanar and bear no particular relation to each other in length. Even though these baselines may not be considered explicitly in signal processing, the system possesses the advantages and the disadvantages of a long baseline system. If ranging stations are separated by some hundreds of miles, making the baselines of the same order of magnitude as the altitude of an orbiting satellite, then the position of the satellite can be located accurately by simultaneous but independent range measurements. In this case, the method of location is that of triangulation, and the uncertainty in position will be that of the individual range measurements, provided that the timing of range measurements can be related to within some figure significantly less than 0.4 msec (which is the ratio of range resolution to velocity for 10-ft accuracy and a velocity of 25,000 ft/sec).

However, when vehicles are being tracked at lunar distances the character of the system changes. The system has perforce become a short-baseline system. As presently envisioned, the system would continue making independent range measurements. The differences between ranges from pairs of stations would define angles, but angular accuracy could be reduced. A range difference would be the difference between two large numbers whose uncertainties increase with range, although it will not be noted that a part of these uncertainties could be made to cancel.

In order to compare the angular resolution of a short baseline system with that of an ultra-long baseline system, consider the change in angle corresponding to a change in range or range difference that equals the uncertainty in range measurement. Consider horizontal baselines, denoting the angle from the vertical by θ. Thus, from

$$\sin\theta = \frac{l}{d}$$

we have

$$\Delta \sin\theta = \frac{\Delta l_1}{d_1} = \frac{\Delta l_2}{d_2} ,$$

where d_1 = length of short baseline system (= 500 ft), d_2 = length of long baseline system, Δl_1 = uncertainty of range-difference measurement in the short baseline system, and $\Delta l_2 = \sqrt{2}$ times the uncertainty of a single range measurement in the long baseline system assuming completely random relative behavior of errors. If f = 400 Mc and if the measurable phase difference $\Delta\phi$ = 2 degrees (which is the accuracy of the present Minitrack system), then

$$\Delta\phi = \frac{2\pi d_1}{\lambda} \Delta \sin\theta = \frac{2\pi d_1}{\lambda} \cdot \frac{\Delta l_2}{d_2} ,$$

whence

$$\frac{d_2}{\Delta l_2} = \frac{2\pi d_1}{\lambda \Delta\phi} = \frac{2\pi \times 500 \times 4 \times 10^8 \times 180}{9.821 \times 10^8 \times 2 \times \pi} = 36,540 .$$

If range can be measured to ± 10 ft then $\Delta l_2 = 10\sqrt{2}$ ft and d_2 = 517,000 ft ≈ 100 miles.

A baseline of 100 miles will provide the same angular resolution from simultaneous range measurements as will a 500 ft baseline system that measures phase to 2 degrees. The angular resolution considered here is 0.03 mrad near the vertical. An angular resolution of 0.01 mrad (10 ft at 200 mi) is desirable and would require a 300-mi baseline. It should be noted that these baselines will give the required precision only if range can be measured to ± 10 ft and if the baselines are laid out with negligible error. At extreme ranges where range accuracy is limited by knowledge of the velocity of propagation, range can be measured only to 1 part in 10^6. Thus at lunar distances (250,000 mi), range might be measured to 0.25 mi accuracy, but if the precision of the measurements is maintained at ± 10 ft the propagation velocity uncertainty will affect each range measurement proportionately and will essentially cancel when the difference is taken. In this case the angular resolution will remain at 0.01 mrad for a 300 mi baseline.

The precision of a short-baseline system that measures range difference directly is independent of range. Thus, it appears that once the satellite range becomes large enough to permit the assumption that plane waves are being received from a distant source, a measuring system that makes independent range measurements at each end of a baseline and computes their difference is inferior to one that measures the difference directly.

If some form of short baseline system is to be used to gain angular precision at long ranges, consideration must also be given to the resolution of ambiguities. Systems such as Minitrack use a series of baselines of decreasing length to resolve angular ambiguities, in effect fixing frequency and varying the baseline. Ambiguities could also be resolved by keeping the baseline fixed and varying the frequency. This could be accomplished by transmitting a carrier and sidetones. The highest-frequency sidetones might be used for angular ambiguity resolution as well as for ranging precision. If a baseline of 2000 ft were used in conjunction with a 400 Mc carrier and a 500 kc upper sidetone, unambiguous angle measurement could be made if phase could be measured to 1 percent and if there were one effective sidetone at 10 or 12 Mc. This effective sidetone might be the difference frequency between the carrier and a multiple of the satellite transponder local oscillator.

It is of interest to investigate the sensitivity to angular change and to frequency change of a short-baseline angle measuring system. Let

$\Delta\phi$ = phase difference between signals at two antennas (horizontal baseline),

θ = angle from the vertical

λ = wavelength of signal from vehicle.

Then

$$\Delta\phi = \frac{2\pi d}{\lambda} \sin\theta \quad ,$$

and the sensitivity to angular change is given by

$$\frac{\partial(\Delta\phi)}{\partial\theta} = \frac{2\pi d}{\lambda} \cos\theta \quad .$$

Thus, the sensitivity is highest near the vertical where signal strength is also greatest.

Now let

$\delta\phi$ = minimum detectable phase change,

$\delta\theta$ = required angular resolution,

$\frac{\delta\phi}{\delta\theta}$ = sensitivity = $\frac{2\pi d}{\lambda} \cos\theta$.

Then, sensitivity to frequency shift or drift is given by

$$\frac{\partial(\Delta\phi)}{\partial f} = \frac{2\pi d}{c} \sin\theta \quad .$$

Thus, with δf the frequency shift,

$$\frac{\delta\phi}{\delta f} = \text{sensitivity} = \frac{2\pi d}{c} \sin\theta \quad ,$$

from which

$$\delta f = \frac{\delta\phi \cdot c}{2\pi d \sin\theta} = \frac{2\pi df}{c}\,\delta\theta\,\frac{c}{2\pi d \sin\theta}\cos\theta = f\,\delta\theta\cot\theta\,,$$

and required frequency precision is

$$\frac{\delta f}{f} = \delta\theta \cot\theta\,.$$

As an example, let $\theta =$ 60 degrees and $\delta\theta = 10^{-5}$ rad. Then $\delta f/f = 5.77 \times 10^{-6}$, and for the conditions specified a frequency shift or drift greater than 5 parts in 10^6 would appear as a measurable error. For satellite orbits, Doppler shifts of 5 parts in 10^5 are expected, this requires that these shifts be allowed for in the phase comparator.

Some investigation has also been made of angle measuring techniques as used by radio astronomers. They have applied the interferometer antenna principle to a variety of methods of making angle measurements. Their techniques invariably result in a measurement of received power variations as the received signals at two antennas alternately cancel and reinforce. The very nature of this power measurement promises crude results as compared with the precise phase measurement of a Minitrack-type system. The precision of angle measurements attainable by this power-measurement practice is considerably less than present satellite tracking system requirements.

Although many of the radio astronomy methods are not immediately applicable to satellite tracking, some techniques may prove useful for tracking lunar probes. One interesting system for orbiting satellites is the "post-detection correlation" interferometer. This technique is intended for use with long baselines, and requires high signal-to-noise ratios. It uses a separate receiver and detector at each antenna of the interferometer. If several frequencies from the same source are received by each antenna, amplified, and detected in a square-law detector, the output of each detector will be the difference frequencies between the signal components, and the relative phases between corresponding difference frequencies at each antenna will give a measure of the path difference to the two antennas. The proponents of this system claim that the two signals will now be of such a low frequency that additional phase shifts introduced by passage through slightly unequal lengths of cable to a common processing area will be negligible. That this claim is fallacious can be seen from the following computation.

Let Δl be the uncertainty in the equality of cable lengths between the antenna positions and the common processing point. This uncertainty may result from originally inaccurate measurement of cable lengths, or from unequal length variations caused by environmental conditions. It will be a fixed error (and hence its electrical effects may be zeroed out in advance for extended periods) only if care is taken to ensure uniform environmental conditions for the cable.

A phase error $\epsilon_1 = k_1 f_0 \Delta l$ will arise from the unequal cable lengths if a single frequency f_0 if received and the phases of the received signals are compared. If f_0 is received with phase ϕ_0 at one antenna and $\phi_0 + \delta$ at the other, the ratio

$$E_1 = \frac{\epsilon_1}{\delta} = \frac{k_1 f_0 \, \Delta l}{\delta}$$

will represent a fractional measurement error.

Now let two frequencies, f_0 and $(f_0 + f_1)$, be received and post-detection correlation be employed. If f_0 and $(f_0 + f_1)$ are from a common source and originate in phase at some time, and if the first antenna receives f_0 with phase ϕ_0, then it will receive $(f_0 + f_1)$ with phase $\theta_0(1 + f_1/f_0)$. The second antenna will receive f_0 at $(\phi_0 + \delta)$ and $(f_0 + f_1)$ at $(\phi_0 + \delta)$ $(1 + f_1/f_0)$. The output of detector 1 will be f_1 at $\phi_0 \, f_1/f_0$ and that of detector 2 will be f_1 at $(\phi_0 + \delta)$ f_1/f_0 . Now, when the signals are sent to the processing area, the unequal cable lengths will contribute phase error $\epsilon_2 = k_1 f_1 \Delta l$. If $f_1 << f_0$, the phase error ϵ_2 will be $<< \epsilon_1$. Note, however, that the fractional measurement error $E_2 = E_1$, that is

$$E_2 = \frac{\epsilon_2}{\delta f_1/f_0} = \frac{k_1 f \Delta l}{\delta f_1/f_0} = \frac{k_1 f_0 \Delta l}{\delta} = E_1 \, .$$

The percent measurement errors are the same in each case, indicating the fallacy of the claim cited above for the "post-detection correlation" interferometer.

Phase-Locked Loops

Generally, a phase-locked loop excited by a signal plus noise can simulate a very narrow-band filter. It actually performs a cross-correlation between the incoming signal plus noise, and a locally generated semblance of the signal. Input noise components whose frequency differences from the incoming signal frequency exceed a small multiple of the nominal bandwidth of the low-pass filter in the loop do not affect the locked oscillation. This statement applies regardless of the nature of these noise components as long as the oscillation is locked to the incoming signal frequency to within a small static phase deviation from orthogonality with the desired input sinusoid. In thinking of the loop as a narrow-band filter, the output noise bandwidth can be shown to be a function of the loop filter bandwidth modified by the input signal level. The modification factor ranges between one and approximately two. The wider bandwidth occurs for large input signals. Therefore, the bandwidth could be made roughly proportional to input signal-to-noise ratio if the rms noise is constant.

Phase-locked oscillators are particularly suited to the function of isolating a desired sinusoid in the presence of other sinusoids whose frequencies, as just noted, differ from the frequency of the desired sinusoid by more than a small multiple of the low-pass loop-filter nominal bandwidth. If the frequency of the incoming sinusoid is exactly equal to the free-running oscillator frequency, then the oscillation will lock to the incoming sinusoid and maintain a 90 degree phase difference with it. Any difference in frequency between the input sinusoid and the free-running oscillator frequency will give rise to a nonzero phase deviation from the 90 degree phase difference. In application to the isolation of the desired sidetones, it is important that no variant and intolerable phase deviations from the ideal 90 degree phase lock exist. Such errors would reduce the potential accuracy of the time measurements that are used in ranging. Sources of such errors in the phase-locked oscillators are either in voltage-controlled oscillator instabilities or in inadequate phase detector circuits or both. In the present application to the isolation of sidetones, phase-detector circuits are not critical because only a small portion of their phase detection characteristics is used. It is of interest here to discuss some of the problems associated with the design of phase-locked loops.

Phase Detectors

Basically, there are two types of phase detectors: the gating type and the diode bridge type. The first type uses the zero crossing of one signal to turn on a current, and the zero crossing of the other to turn off that current. The output is then integrated to give a voltage proportional to the phase difference. If both signals are thought of as square waves, the use of a logical AND, or coincidence, circuit produces an output only when both signals are, say, positive. When integrated, the output is triangular rising from 0 at phase opposition to a maximum value when in phase. Since the circuit gates a current, the output contains a dc component (the mean value of the wave).

If the square waves are differentiated, and the positive pulses from one signal gate on a current, while the positive pulses of the other signal gate off the current, a linear phase progression produces a linear output from 0 to 360 degrees; but here also, a dc component is present in the output wave.

The balanced diode bridge (Figure A1), or product detector, will produce excellent results as a phase detector. Its operation is similar to that of the logical AND circuit but, usually, sine waves are used for input signals. In this case, the linear phase progression produces a *zero-mean* sine wave output with zero crossings at 90 degrees and 270 degrees phase difference. If square waves are used for input, the output is a zero-mean triangular wave with zero crossings at 90 degrees and 270 degrees input phase difference.

A major point in favor of the use of diode-bridge type phase detectors is the fact that the output has a zero mean that does not vary with signal level or noise bandwidth. With sine wave inputs, however, the slope at zero crossing will change with a change in signal level.

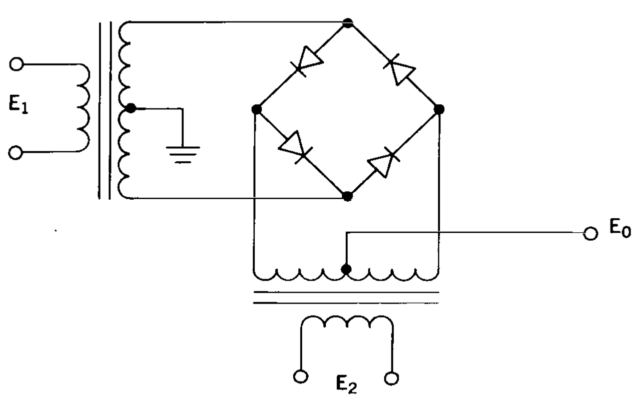

Figure A1—Diode bridge phase detector.

Since the intended use of the phase-locked loop requires very small phase errors, the use of the diode-bridge type detector is not only desirable, but nearly mandatory. The errors in phase shift through the detector must be much smaller than 1 percent and, therefore, the operating point will be within 3.6 degrees of 90 degrees. Since this is such a small angle, the sinusoidal output of the diode bridge phase detector is very nearly linear over the required range.

Voltage-Controlled Oscillators

The width of the spectrum of the output of a free-running oscillator is a measure of the coherence of the socillation. Slow drifts in the frequency location of this spectrum are readily corrected for in a phase-locked loop. But the disturbances represented by the spectral components that are displaced more than (nearly) the nominal bandwidth of the open loop frequency characteristic from the center of the oscillation spectrum cannot be compensated for by the action of the loop. (This action simulates a high-pass response to internal variations.) Unfortunately, an oscillation whose spectrum is extremely narrow does not yield easily to voltage control over a wide range. Therefore, some sacrifice of oscillator coherence (spectral narrowness or short-term stability) must be made in order to achieve the desired control over the center frequency of the oscillation. The usual approach is to use a crystal oscillator whose short-term stability is very good, and use some variable reactance (which tends to degrade this stability) to control the frequency of oscillation. It is doubtful if the multivibrator-type VCO would be stable enough to produce a satisfactorily usable system, because of the inherent voltage sensitivity of the circuit using semiconductors.

Loop Filter

A considerable amount of work has been done and published on loop filters. In a noisy environment, a phase-locked loop will work best with an optimized lag network. Since this will give us the minimum mean square error in the output signal, it is the type that should be used for this system.

Zero Phase Error

Since an error voltage from the phase detector is necessary to make the VCO oscillate at a frequency other than its free-running frequency, a phase error must occur in the steady-state condition. The only time zero error voltage is fed back is when the input signal is at the natural oscillating frequency of the VCO.

In the system required to provide range information, the phase error must be small to stay within the required range accuracy. There are two ways to make the error voltage come out zero, one is to use an infinite gain amplifier in the error loop, and the other is to provide a method of making the error voltage accumulate and be self-zeroing. Since the error must accumulate, a time constant is necessarily involved and zero error can occur only after infinite time. It would seem that the time lag would provide a better solution than the high gain.

Zero-Error Extraction of a 100 kc Sidetone

The system illustrated in Figure A2 is proposed here for extracting a 100-kc sidetone with a zero phase error. The operation may be explained as follows. Loop 1 is a simple phase-locked loop, and as the 100 kc moves an amount δ from VCO-1 natural frequency, the loop error pulls off zero to make VCO-1 fall exactly on the frequency of the 100 kc + δ input in the steady state. This voltage is proportional (by the VCO-1 transfer characteristic) to the frequency offset of the 100 kc + δ

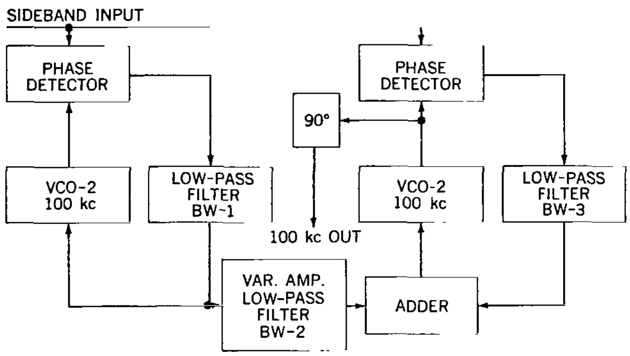

Figure A2—Zero phase error extraction of a 100 kc sidetone.

input. We may make use of this voltage to move another identical VCO-2 to the frequency of the incoming 100 kc signal.

In the steady state, VCO-2 and VCO-1 will both have exactly the same frequency as the 100 kc + δ input signal even though loop 2 is not closed. If we close that loop temporarily, the loop will pull VCO-2 directly 90° out of phase with the 100 kc + δ input, and the error meter will decay to zero as $t \longrightarrow \infty$.

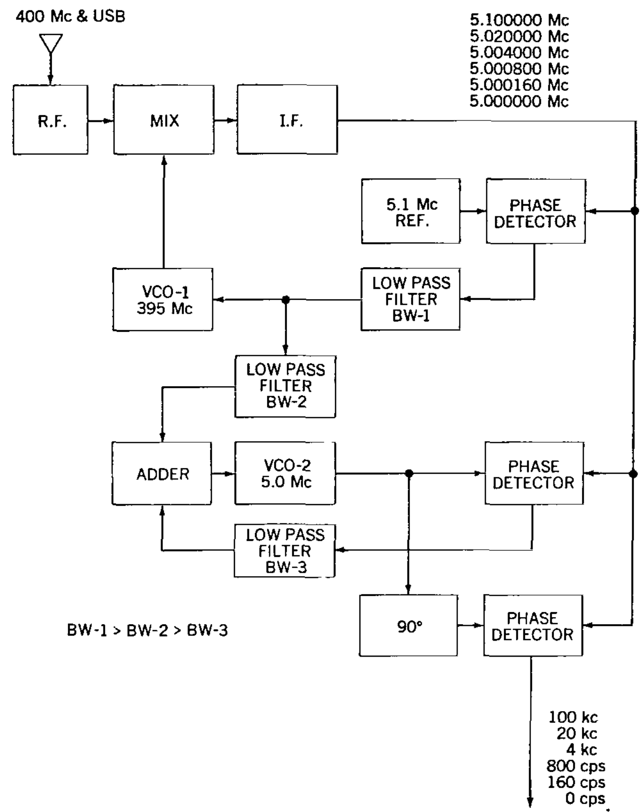

Figure A3—Small phase error sidetone extraction receiver.

The above will be true under the following conditions:

1) Steady state $(t \longrightarrow \infty)$, and

2) VCO-1 transfer characteristic identical to VCO-2 transfer characteristic.

The system is essentially an open loop correction on the VCO-2 frequency, and the error may be adjusted, by the gain balance, from positive through zero, to negative.

There are many methods to instrument this type of a system. One is to obtain the Doppler information from the tracking receiver VCO in the manner illustrated in Figure A3. With reference to this figure, we note that ϕ_e will go to zero throughout the operating range if:

1) The VCO-1, zero-error point has been moved to the Doppler zero point;

2) The matching between the VCO-2 transfer characteristic and the VCO-1 transfer characteristic holds over the operating range;

3) The phase shift in the detection loop is exactly 90 degrees;

4) The antenna and RF stage are phase linear around 400 Mc ± 10 kc;

5) The IF strip is phase linear around 4.9 Mc ± 2.5 cps.

Another application of the same technique is shown in Figure A4. An investigation of this system shows that with no signal input, the VCO will search for a signal within limits provided by the discriminator when the discriminator loop gain is greater than unity, will lock on a signal when one comes in, and will have a steady-state phase error of very nearly zero. In order to make the conditions such that sweep search can take place, the discriminator loop gain must be greater than unity (open loop), and, therefore, any phase error present will be negative, but arbitrarily small.

Critique of a System Under Consideration by Goddard Space Flight Center

The system still under consideration by GSFC at this writing provides a 100 kc synchronous detector as illustrated in Figure A5.

As stated earlier, a diode-bridge type phase detector would be superior to the coincidence type phase detector used by GSFC.

Again in the light of our introductory remarks on phase-locked loops, the 100 kc filter that precedes the phase-locked loop is superfluous if it is only intended to eliminate the other sidetones. A phase-locked oscillator whose low-pass loop bandwidth is of the order of a few cycles will, if tuned to the desired input sidetone, extract this sidetone within the desired phase accuracy as long as the nearest sidetone to it is a few tens of cycles or more away, assuming that the background noise can be ignored. If the total background noise is very strong, the same conclusion should hold as long as the free-running oscillator frequency coincides with the frequency of the desired input tone. Otherwise widening the noise bandwidth before the phase-locked loop by a large factor may

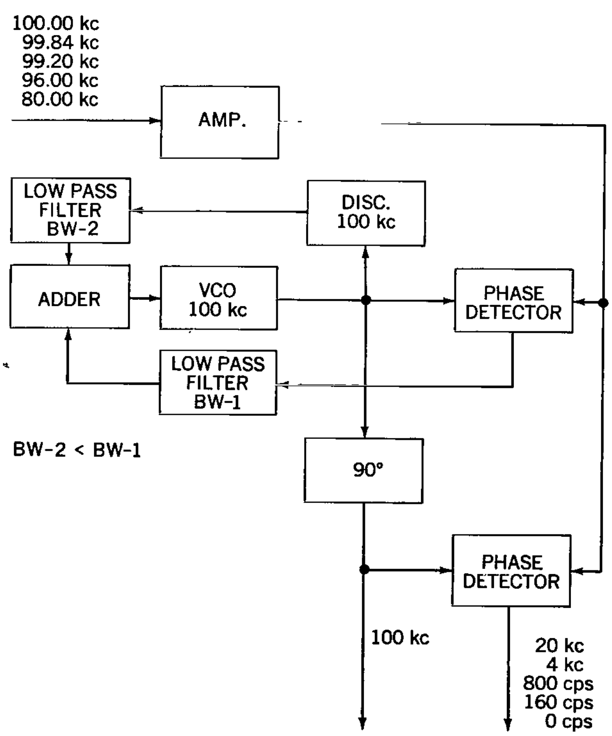

Figure A4—Small phase error phase-locked loop.

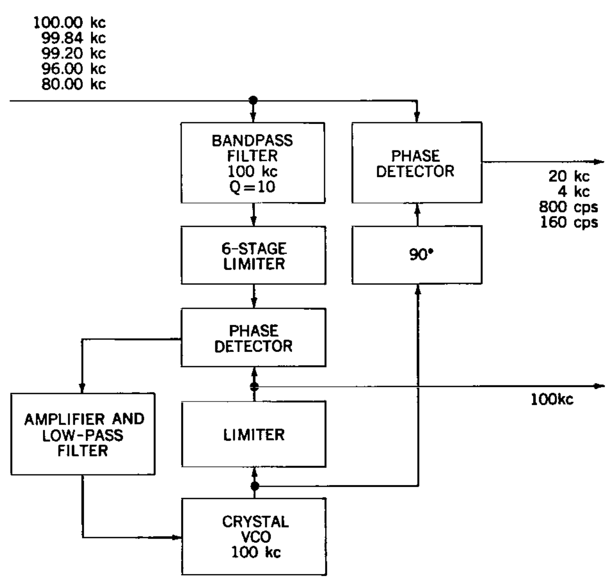

Figure A5—Original phase-locked loop.

cause a large rise in the locking threshold. Amplitude limiting before the phase-locked loop is desirable in combatting random-fluctuation noise, but it is not basically important to the isolation of the desired sidetone from among the other sidetones as long as the various frequency differences are sufficiently high in comparison with the loop bandwidth.

Task I Recommendations

Signaling

In the light of the investigations carried out thus far at ADCOM, five methods of signaling stand out among the many possibilities as possessing, to varying but reasonable degrees, the merits of simplicity, efficiency, capability to yield satisfactory accuracy, and potentiality for high immunity to many important varieties of noise and disturbances. These are:

(1) SSB modulation of a carrier by a set of appropriately chosen sidetones. The high-frequency sidetones should be favored with the high amplitudes, and the phases of the returned lower-frequency sidetones should be favored with the longer integration times.

(2) Frequency modulation of a carrier by a set of appropriately chosen sidetones, phase demodulation of the FM signal returned by the transponder, followed by isolation of the desired tones by means of phase-locked loops whose oscillations are then used directly (without any further phase shifting) for phase comparison operations leading to the extraction of the desired ranging information. The sidetone amplitudes fed to the ground FM modulator must be so pre-emphasized that after two integrations they emerge with the higher-frequency sidetones still more heavily favored than the lower-frequency sidetones. Again the longer integration times should be used in processing the phases of the lower-frequency components.

(3) Frequency or phase modulate a carrier by one or at most two appropriately chosen sidetones, and isolate a set of desired sidetones directly from the spectrum of the returned signal (without a prior use of an exponent demodulator) by means of phase-locked oscillators.

(4) A combined technique that employs one of the above signaling methods until the tracked object becomes so distant that the returned signal falls to some pre-specified design level, then the signaling tactic is switched to a frequency-hopping mode with one or perhaps no more than two tones being transmitted at a time.

(5) Use a single carrier with rounded-envelope phase-reversal keying according to some carefully planned binary coding scheme that embodies a combined ambiguity-resolving as well as high-accuracy ranging code groupings. A closely related alternate might use two phase-reversal keyed carriers—one intended for coarse resolution, the other for fine measurement. The "baud" length of the coarse-resolution carrier is varied in a discrete manner according to current data on the range so that it will automatically continue to provide the necessary "coarse" resolution as the object recedes deeper and deeper into space.

The only merit that can be cited for SSB over the others is its bandwidth economy. For tracking under adequately favorable signal-to-noise ratios, a more desirable alternative is a combination of techniques (2) and (3) in which the ground transmitter uses a few tones (intended principally for coarse ranging) to frequency modulate a carrier with a moderate deviation ratio, and the ground receiver operates on the returned signal with a phase demodulator followed by phase-locked oscillators in order to extract the "coarse-ranging" tones, and in addition operates directly upon the i-f spectrum with phase-locked oscillators to extract high-frequency tones for "finer-ranging" measurements.

Techniques (4) and (5) are particularly suited for deep-space-probe ranging because of the more efficient manner in which they utilize available signal power. Frequency hopping may well prove to be the most desirable signaling for deep-space ranging. However, phase-extrapolating locked-oscillators must be developed for frequency hopping, and requisite coding strategies remain to be devised for phase-reversal keying.

Transponder Design

With FM and unrounded-envelope phase signaling, design an amplitude-limiting class C power-amplifying repeater which translates the incoming spectrum from the neighborhood of one frequency to the neighborhood of another in such a way that transponder-oscillator drifts are automatically subtracted out.

If SSB signal or if rounded-envelope phase-reversal keying is used, a linear repeater must be used with class B power amplification.

A transponder design with "narrow-band" and "wide-band" modes of operation as explained earlier merits serious consideration, especially for deep-space probes.

Phase-Locked Oscillators

It is recommended that crystal oscillators be employed because of their characteristic stability in the absence of external control. The importance of stability stems from the desire to use these oscillators in the simulation of very narrow tone-isolating filters. Internal instabilities may cause phase errors that are not compensated for automatically by action of the loop.

Further improvements of phase-locked oscillators by means of higher-order loop corrections are needed to reduce static phase errors that result from small (Doppler) frequency shifts of incoming tones.

A diode-bridge type of phase detector is also recommended for use in phase-locked loops because it appears to be more capable of producing a stable null operation at center frequency than other alternatives.

Program for Next Interval

The following program is planned for the next interval.

Theoretical Investigations

1. Further study and evaluation of the application of exponential modulation techniques to sidetone ranging.

2. Study of available information on the anomalies of propagation within 250,000 miles of the earth's surface and consideration of pseudo-random signaling techniques as well as frequency hopping and phase-reversal keying in combatting the random effects of propagation and of additive random-fluctuation noise.

3. Implementation of frequency hopping, phase-reversal keying, and other exponential modulation techniques for ranging.

4. Further study of angle and angle rate measurement techniques.

5. Study of Doppler effects upon range and angle measurements and their application to range rate and angle rate measurements.

6. Study of sources of phase errors caused by spurious components in output of envelope detectors, class C amplifiers and amplitude limiters.

Experimental Investigations

1. Studies of stability and maintenance of zero-phase errors in phase-locked loops, in particular, breadboarding of 100-kc zero-error phase-looked loops of the type illustrated in Figures A2, A3, and A4 for the purpose of determining the long-term and short-term stabilities as well as the degree of improved performance resulting from the use of such systems.

2. Comparative study of various types of phase detectors used in phase-locked loops.

3. Development of phase-locked oscillators that are capable of extrapolating the smooth phase behavior of signals of the type used in sidetone ranging, on the basis of observed behavior during some preceding time interval of specified duration.

NASA-Langley, 1964 G-235